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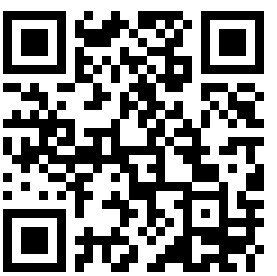

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The Society of Telegraph Engineers

FOUNDED 1871

INCORPORATED BY ROYAL CHARTER 1921

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WIRELESS SECTION: CHAIRMAN'S ADDRESS.*

By Captain C. E. KENNEDY-PURVIS, R.N., Member.

(*Address delivered 6th November*, 1929.)

In this address I shall attempt to deal briefly with the progress of the art of wireless communication during the past two years.

One of the outstanding events is the fusion of wireless telegraph and cable interests by the merging of the communication side of the Marconi Co. with the Eastern Telegraph Co. Opinions as to the advisability of this are various, but the scheme has the approval of the Government and is now an accomplished fact. My own point of view is that this fusion is all to the good. Wireless telegraphy and cables are complementary to one another for long-distance point-to-point communication. It would have been little short of disaster for the cable services to have been completely superseded.

We are in an age of acute competition in communications, and every possible line must pull its pound. Cables have a distinct rôle in this, and after their long period of monopoly their owners are being forced by stress of circumstance to push on with developments and improve their technique. Some of these improvements may make the wireless world feel that cables are still a keen competitor.

We may now look forward to concerted development of Empire and world communications using both systems under one authority.

The Imperial and International Communication Co. has a link with the Governments of the United Kingdom, Dominions and Colonies through an advisory board of nominated members which has wide powers and can control the work of the company to a considerable extent, from the point of view of rates and service generally. There seems little doubt that there is a mass of traffic to be picked up if rates are made sufficiently low, and the cable and wireless technical staffs must so improve working and speed of traffic that rates can be cut and yet still pay. From the strategical point of view, cables are considered to be essential, and the merger saves these for future use in times of emergency or war.

The development of Empire and world communications by wireless telegraphy has shown satisfactory progress during the past two years. The original beam services to the British Dominions have continued to give an excellent and improved service, and the beam routes to South America and the United States are now well in their stride. Subsidiary lines by short wave to the smaller Colonies are being developed, and will develop much more quickly when the merger company has time to formulate its full proposals. Foreign nations are developing their own lines of world-wide communication, and competition among all nations to gather in traffic is acute.

While the use of short-wave working for long-distance traffic is advancing by leaps and bounds, I venture to put in a plea that the development of long-wave working should not be entirely abandoned. Congestion in the ether seems likely to be permanent; congestion will probably always be ahead of technique. As technique improves, additional lines will be started up and will overlap such improvement as may be effected from time to time.

It will be foolish to desert the lower frequencies without exploiting them also to the uttermost. In time of war, even if we are neutral, it may be found that short-wave working becomes exceedingly difficult; a nation at war will not care much for conventions, and those countries which have continued to use a long-wave service and their cables will have an advantage.

Radio telephony has made very great strides and has now reached the stage of being a sound commercial proposition. Commercial telephony over long distances was first made possible by the British Post Office in conjunction with the American Telegraph and Telephone Co., with the result that Great Britain is the centre for Europe of the North American transatlantic service. By the judicious use of long and short waves a 24-hours service is possible, and four separate channels are available.

Other countries have since advanced with more or less success, and some services now opened are Holland to Java; Germany to Siam and Buenos Aires; France

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to Buenos Aires; and Spain to the United States and Buenos Aires.

Experiments in multiplex transmissions whereby one carrier is made to do duty for one or more channels of telegraphy as well as telephony show signs of promise, but it is rather a question whether the wave-band occupied by such methods at present is not considerably wider than the sum of the wave-bands which would be used by an equal number of separate stations.

Cable companies are not lagging behind; a cable capable of telephony and high-speed telegraphy across the Atlantic is likely to be put into service before long.

Turning to broadcasting, the difficulties which arise between different nations as to their requirements are now in a fair way to be considered as settled, at least for a time. There are more than 150 broadcasting stations operating in Europe alone. It is easy to see that without very careful organization there must be serious interference over the whole of Europe. The recent agreement of 27 European nations to the Prague Plan for the allocation of wavelengths was a notable step in the right direction.

This Plan is the culmination of four years of work at the Union Internationale de Radiophonie. As early as 1924 it was realized that international agreement and co-operation were essential if the rapid increase of the number of European stations was not to result in a state of complete chaos. This was due to the fact that the majority operated on wavelengths below 500 m, and that, although the range of the ground wave may be limited, under certain conditions the range of the indirect wave may cause interference at a very much greater distance.

Two international plans preceded the Prague Plan—the Geneva Plan of 1926 and the Brussels Plan of 1928. These two plans were successful in so far that friendly agreements were come to by responsible authorities to operate their stations on certain wavelengths. They suffered from the disadvantage that they were not agreements between the Governments of the various nations, and there was therefore no method of securing adherence to them.

The Prague Plan was a great advance on its predecessors in that it is an agreement between Governments and can thus be enforced. Although the allocation of wavelengths has been much criticized it was obvious that some such plan was necessary, and there had to be a certain amount of give and take. Nations who were early in the field, e.g. Great Britain, France and Germany, suffered somewhat on the re-allocation of wavelengths, but this was in a good cause. The establishment of an international wavelength checking post at Brussels was also a step in the right direction.

One important point which will soon require settlement is the encroachment of long-wave broadcasting on to the wave-band allocated for use by aircraft.

In Great Britain, interest has been aroused amongst the public by the Regional Scheme under which the country is to be served by a series of twin high-power stations located at selected points outside certain cities. The recent opening of the new station at Brookman's Park marks the beginning of this scheme. The underlying principle is to give to every listener an alternative programme, whilst to many listeners with selective sets it will provide several alternative programmes. In addition it will give a maximum of field strength over the larger cities where local interference is caused by trams and other electrical machinery. It carries with it, however, one drawback, viz. a probable "wipe out" area close to the station from which it will be difficult to receive anything but the local transmission and in which the listener may experience difficulty in separating the two local programmes if the simpler types of receiving sets are used.

Communication with ships must always interest a country which depends on ships for its existence. Services to and from ships have steadily improved. High-speed working in the case of big ships of the western ocean trade, and short-wave working for very long-distance communication, have developed. Long-wave working for medium distances has also shown advance.

That bugbear to broadcast listeners—"spark"—is being gradually scotched by the introduction of interrupted-continuous-wave working at all Post Office coastal stations. It remains for other nations to follow suit. An interesting point in this connection is the use of different note frequencies by each British coastal station in order to avoid congestion of traffic, it being a fairly simple matter to over-read or under-read the signal required from the interfering signal.

Let us now consider the theoretical advances which have been made. A very considerable addition has been made to our knowledge of the theory of the propagation of waves, notably by research work carried out in England under the auspices of the Radio Research Board and by the Marconi Co.'s research staff. This work has been described in papers read before the Wireless Section, and the earlier theories of Eccles and Larmor regarding the mechanism of wave propagation have been largely confirmed by the experimental work of Appleton, Eckersley and Hollingworth. The work done has shown that, whilst normally the long waves are plane-polarized, an important change occurs in the medium waves of the higher frequencies which may have downcoming components the polarization of which differs from that for the surface wave. The long waves travel over the surface of the earth, whilst the higher-frequency waves in the medium band not only travel over the earth's surface but have indirect components which are free from the earth and are modified by the Heaviside layer. A combination of the surface and indirect components of the medium waves may produce very serious night errors in direction-finding, and fading due to the interference of these components also occurs.

As regards the short-wave band, very valuable evidence has been obtained on the fading of wireless signals. This has been dealt with very fully in a recent paper read before the Wireless Section. Experimental work indicates that the Heaviside layer is a complex structure of electron clouds or strata, and when a group of electromagnetic waves at the frequencies of the short-wave band enters the layer the following phenomena occur:—

(1) *Scattering of the radiation.* This is evidenced in a remarkable manner by the fact that, for the case of the Marconi short-wave beam transmitters working on

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a receiving station, direction-finding observations taken at distances close to the transmitter and within the skip zone give the bearing of the place where the beam enters the Heaviside layer, which corresponds to the great-circle bearing of the receiving station.

(2) *Fading of signals.* This arises from the complexity of the layer and the fact that, in addition to their purely free irregular motions, the electrons may also have definite motions due to the magnetic field of the earth. A plane-polarized wave entering the layer must ultimately be resolved into at least two circularly polarized waves travelling with different velocities, whilst for the case of a group of waves such as go to make up a signal, different parts of the group will also have different velocities. The various fading effects may be considered to be due to:—

(i) *Rotation* of the plane of polarization of the combined result of these circularly polarized waves, due to slight changes in the magnetic field.

(ii) *Interferences.* There seems to be experimental evidence that the waves reaching a receiving station may have traversed paths of very different lengths. These can interfere with one another and thus produce fading. Some of these have suffered only a single reflection in the Heaviside layer, while others have suffered multiple reflections between the Heaviside layer and the earth. The term "reflection" is here used in a very general sense to indicate that the wave has been bent back to the earth by the Heaviside layer. Calculation shows that modulation of the transmitted waves should reduce fading from this cause, and recent tests indicate that this is actually the case.

(iii) *Scattering.* When scattering is present, further fading effects are thereby caused.

The methods employed in short-wave reception for mitigation of the effects of this signal fading include the use of spaced aerial arrays and long horizontal aerials of the Beverage type. In addition, circuits have been developed with a view to obtaining increase of sensitivity for weaker signals, e.g. the use of grid bias the value of which depends on the field intensity at the receiving aerial.

The research work and experimental observations which have been made on the Heaviside layer, and on which papers have also been read before the Wireless Section, have indicated that there are two or more regions where changes occur, the two main regions being at heights of the order of 100 and 250 km respectively.

The problems underlying this work have been very largely tackled by the employment of suitable signal intensity and direction-finding apparatus which has also been described before the Wireless Section, and this work has resulted especially in a very considerable development of aerials and apparatus suitable for short-wave direction-finding, these aerials being based mainly on the type originally proposed by Adcock, in which the receiving aerial system is so designed that it is unaffected by any horizontal component of electrical force.

As regards the possibility of taking direction-finding bearings on short waves, valuable evidence illustrating the limitations has been published. The direction-finding observations which have been made on distant and near stations have tended to confirm the general correctness of our ideas regarding the Heaviside layer and wave propagation therein. Summarizing the results from these observations the following broad conclusions are obtained:—

(1) It is reasonably possible to locate by direction-finding apparatus a very near-by transmitting station where the ground wave preponderates.

(2) If the distance is such that the point of observation is within or close to the edge of the skip distance then the whole of the effects are due to scattering and either little or nil bearing, or an enormous error, is observed.

(3) As regards the distant transmitting stations, (*a*) If the distance is so great that in general a single indirect ray only is being received, bearings are reasonably accurate. (*b*) Even at very great distances, however, though bearings definitely in error are not frequent, there are some occasions on which nil or indistinct bearings are obtained, owing doubtless to either (i) a multiplicity of waves following different paths, or (ii) interference arising from scattering.

Our knowledge of the echo effects produced at a receiving station on the shorter waves (which may give rise to mutilation of signals) has been increased, and records have been taken with the cathode-ray oscillograph showing the arrival of waves which have travelled several times around the earth's perimeter.

A very interesting phenomenon is the arrival of echoes after a very long interval of time, and intervals of 15 seconds have been observed by continental research workers.

Whilst some workers are of opinion that these echoes are due to special conditions in the Heaviside layer enabling the waves to pass many times around the earth, there is no adequate explanation at the present time of how the great attenuation that one would expect in such cases is overcome. An alternative explanation is that these waves have penetrated the earth's atmosphere and have been reflected back by swarms or bands of electrons in space. This view has been put forward by C. Störmer.

The echo phenomenon observed in short-wave reception has led to the development of directive aerial systems tending to reduce the mutilation which the echoes otherwise cause.

Taking next the development of the use of long and medium waves, the valve thermionic generator is continuously displacing the other and older systems of high-frequency generation. It is only on the longer waves that the high-frequency alternator and arc systems are still being employed for point-to-point services, and the tendency is to reduce the traffic hours of these long-wave systems by supplementary valve short-wave systems.

Advance has been made in the way of stabilizing emissions, which in these days of congestion is a necessity.

As regards broadcasting, an enormous amount of hard technical work has been necessary to produce the extremely good quality of the B.B.C. transmissions.

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I have mentioned the Regional Scheme earlier in this address. From the technical point of view, perhaps the most interesting development which the regional scheme renders necessary is the placing of all the so-called relay stations on the same wavelength. This proposal has brought with it the great problem of obtaining isochronism of the carrier wave of all the stations, and the success of the scheme must depend largely on the solution of this problem. So far it has been attacked by the use of tuning forks as the fundamental standard from which the carrier waves are built up by frequency multiplication. Apart from the disadvantage of the complicated amplifier system for frequency step-up, this method appears to give the required degree of accuracy. The advance being made in the technique of quartz-crystal control may ultimately, however, offer an alternative solution.

Short-wave development has been very rapid, whilst the short-wave beam and similar services have been well maintained. The main development of short waves has been in regard to the great extension of the use of waves in the higher-frequency regions between 15 000 and 21 000 kilocycles per sec., which band of frequencies has been found effective for long-distance communication during daylight hours.

Still higher frequencies have been used with a certain amount of success but, in general, it has been found that the results obtainable at frequencies from 21 000 to 30 000 kilocycles per sec. are subject to very heavy fading. During the cruise of H.M.S. "Yarmouth" to the Far East, commencing December 1925, good communication was occasionally established with Horsea as far as Hong-Kong, using a frequency of 25 000 kilocycles per sec. The Marconi Co. have obtained results across the Atlantic, using a wave of 8·67 m (34 600 kilocycles per sec.).

In regard to the frequency band between 15 000 and 21 000 kilocycles per sec., the results obtained have shown that the hours of working vary considerably with the season, but the development of these waves has enabled communication to be obtained at long ranges over a very large proportion of the 24 hours by the judicious change of frequencies at various times of the year. The success obtained with this frequency band is due not only to improvements in transmitting circuits and valves, but also in considerable measure to the adoption of special forms of directive aerial systems which have been found very efficient for radiation at these frequencies.

Apart from the Franklin vertical arrays as used by the beam stations, the most generally used types for these frequencies are perhaps multiple arrays of horizontal or vertical dipoles, but some interesting systems have also been developed in France by Mesny and Chiveix in which the array comprises wires arranged to form either a double zigzag or a Grecian Key curtain, and for long-range working multiple arrays erected on low masts of height less than 150 ft. have given satisfactory results.

Commercially, as regards the type of aerial array for a given service, it seems to be a problem of balancing power input and its cost against the cost of masts.

The Marconi Co. have developed their beam system for operation on this higher-frequency band, and considerable work has also been done on the Continent. For example, a service with South America is maintained by Germany, and the Dutch Government has developed a very considerable communication service between Holland and Java, using frequencies of the order of 18 000 kilocycles per sec. for day work and 10 000 for night work, variations in frequency being made from time to time as found necessary owing to the seasonal changes.

During the European summer period, particularly about June, practically continuous communication is maintained throughout the 24 hours of the day using these higher frequencies only. In winter time the higher-frequency communication is limited to some 4 hours of the day, but the lower-frequency period can be carried on for approximately 12 hours. For the communication by Germany with South America, frequencies of the order of 16 500 and 11 500 kilocycles per sec. are employed.

The Admiralty have for some years been working a service to the East (Hong-Kong and Singapore) on 8 570 kilocycles per sec. This service has now been supplemented by a day wave service working on about 18 000 kilocycles per sec., and during June of this year a 24-hour service was maintained with Singapore on the latter frequency. During the winter months, however, the possible service is much reduced, but the daily hours of communication can then be augmented by the use of frequencies of the order of 13 000 kilocycles per sec.

Other interesting applications of short-wave working have been the use by the British Post Office of short waves of the order of 19 000 kilocycles per sec. for the transatlantic telephone system as an alternative to the long-wave transmission, and the short-wave broadcasting programme on 12 500 kilocycles per sec. which has been carried out from Chelmsford. In addition, in Holland a short-wave transmitter has been installed near Hilversum with the object of broadcasting direct to the Dutch East Indies on the daylight wave-band.

All the above developments refer to point-to-point or to broadcasting services, but there has been a very considerable development in the use of short waves for mobile services both for naval working and for the mercantile marine. As regards British naval services, practically all H.M. ships have been adapted for short-wave working.

Satisfactory results have been invariably obtained on the lower frequencies usually employed for the night services, but it is not so easy to obtain satisfactory results on the daylight wave-band and the difficulties are at present not entirely overcome. Although the use of short waves for mobile ship services is proving of great value, there is a certain limitation to its use, due chiefly to the zone where either interference is experienced between the surface and indirect radiation from the transmitter, or there is a definite gap between these components of the transmission.

This "skip zone" has been observed at sea on all frequencies above 6 000 kilocycles per sec. at varying distances below 1 000 miles. The lower-frequency band between 6 000 and 4 000 kilocycles per sec. has been

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found to be very much better in this respect, but it cannot be said that short waves will meet the requirement of a continuous service for ranges up to 1 000 miles at sea to the same degree of reliability as long waves.

Apparatus Development.

As regards the obtaining of H.T. power supply for transmitters, the high-tension d.c. generators for the Rugby long-wave high-power station have now been in operation for some years and have given very satisfactory results. The British Broadcasting Corporation are also using high-tension d.c. power, for example in their 5GB station, with satisfactory results.

The alternative systems of H.T. supply by means of single-phase or multi-phase rectifier systems are, however, in more general use for the medium-power transmitters, and especially is the use of double 3-phase rectification in favour. It has been claimed that the rectifier systems give greater security and reduced tendency to flash-over of valves. On the other hand, figures have been published showing a considerable economy resulting from the use of high-tension d.c. machines for fixed station transmitters.

For the high-power medium- and long-wave transmitters, the metal-glass valve with suitable cooling agents, such as distilled water or oil, has become well established; the silica radiation valve has also shown itself to be a very economical means of generating oscillatory power up to 10 and 20 kW, especially where the space is limited. Considerable experimental tests on metal-glass valves of increased anode rating have been carried out by the Post Office engineering staff at Rugby, but the development of higher-power standardized valve units capable of being used in parallel is proceeding more slowly than at one time appeared probable.

The development of the daylight wave-band of frequencies for short-wave service working on long ranges has increased considerably the technical problems connected with transmitting valves and circuits. The inherent capacity of the valves imposes a limit to the permissible inductance and, in practice, it is a matter of extreme difficulty to design stabilized transmitters for these very high frequencies. As regards the metal-glass valves, there is an increased danger of puncture of the glass envelope; this is usually guarded against by fitting an annular ring as an internal guard-ring device to screen the seal joint of the metal to the glass envelope.

In this connection it may be of interest to record that, as regards high-power valves used for Admiralty services on these frequencies, very considerable success has attended the use of the water-cooled silica valve the development of which has been described before the Wireless Section, and two of these valves are in regular operation handling from 20 to 30 kW for Admiralty services to the Far East on frequencies of the order of 18 000 kilocycles per sec.

The ability of the silica envelope to withstand very high voltages at frequencies as high as 30 000 kilocycles per sec., without danger of puncture, renders it peculiarly suitable for the generation of oscillatory power on these higher frequencies.

The increase in wireless telegraph services is bringing with it an ever-increasing tendency to greater stringency as regards frequency control of transmitters. The tuning-fork method of control has now been in use for some considerable time in this country, e.g. at the Rugby high-power long-wave station and at various stations of the British Broadcasting Corporation, and this has proved an excellent system of high constancy.

As regards short-wave transmitters, the very large number of stages which would be required for tuning-fork control of the short-wave band of frequencies is naturally a disadvantage and, for this reason, the short-wave transmitters are controlled in frequency either by well-designed short-wave master control circuits as used by the beam stations, or by the use of quartz crystals. The quartz crystals employed have usually a fundamental frequency of the order of 2 500–3 000 kilocycles per sec., and stabilized transmitters taking H.T. inputs of 30 to 50 kW at their final stage are in operation using frequencies up to the sixth harmonic of the quartz crystal.

The use of quartz may perhaps not give such a high constancy of frequency as that which is obtained from tuning-fork control and, owing to the number of modes of vibration, care has to be taken to ensure that maximum control is obtained on the correct frequency. It does, however, operate exceedingly satisfactorily when the amplifying circuits are properly tuned and neutrodyned, and is capable of giving a continuous-wave note of excellent quality even for the case of relatively poor regulation of the H.T. and filament power supplies for the transmitter, or when a.c. supplies are used for the filaments.

There is usually a change of frequency with temperature of the order of 30 parts in 1 million per degree C. with the use of quartz crystals, and for this reason thermostatic controls are used, but it has been claimed by certain experimentalists that careful selection of the quartz cuts may reduce this variation to the order of 2 parts in 1 million.

Turning to the development of receiving apparatus, there has been very marked advance in connection with the design of broadcasting receivers and their components, and the apparatus sold by the great majority of British firms is of excellent quality and most reliable in operation.

There has been a gradual change in the type of set offered to the listening public by the manufacturers, consequent on the introduction of two new valves to the market, i.e. the screen-grid valve and the pentode. The large amplification factor of the former and the large mutual conductance of the latter have made it possible to build up relatively cheap sets using only three valves and having a performance at least equal to that of the more ordinary types of sets employing four 3-electrode valves. From an operating point of view these sets gain in ease of handling, but they need careful construction and screening in order to function properly.

The moving-coil loud-speaker has also made great progress, and there has been a general development of power amplifying valves which has resulted in greater demands being made upon the high-tension supply of broadcast sets. This has led to improvements in dry

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In the case of mobile stations the right to use any wave of the bands applicable was agreed to.

It was agreed that each nation should establish a national laboratory with a national frequency standard, each country to monitor frequencies of its own stations.

In this country the N.P.L. standard will be regarded as the normal standard, and the Post Office will do the necessary monitoring.

A " pious hope " was expressed that countries would hasten the abolition of spark, even before the date fixed by the Washington Convention.

Questions (7), (8) and (14) dealt with the channelling problem. The channelling scheme itself was not agreed to, but the Committee drew up certain recommendations necessary in order to prevent chaos in allocation of wavelengths. These briefly are:—

(1) In the reserved fixed service bands below 50 m, frequencies allotted in kilocycles should be divisible by 5. It was agreed that a separation of 5 kilocycles is not always sufficient at present but that actual practice shows a separation of 0·1 per cent to be practicable.

(2) Certain definite particulars should be forwarded when notifying waves. This was intended to stop premature wave-appropriation, but whether it will do so or not remains to be seen.

(3) Indicates the frequency band width (in cycles) occupied approximately by various forms of wireless transmission.

(4) Stresses the necessity of selective reception.

(5) Deals with the question of the use of short waves, of different orders for various distances. Final recommendations were:—

(*a*) Waves between 50 and 13 m should be reserved for long distances, but these may be used for moderate distances from 2 hours after sunrise until 2 hours before sunset.

(*b*) Waves between 50 and 85 m would be reserved for moderate distances.

(*c*) Waves between 85 and 200 m would be used for the shorter distances.

These recommendations refer to point-to-point working only, and are only put forward as a general principle.

(6) Draws attention to the importance of stations which use the shared bands employing modern apparatus, having regard to the number of stations which are likely in the near future to operate in these bands.

Question (15) dealt with the limitation of power of broadcasting stations, and it was recommended that the power of European broadcasting stations using waves between 200 and 545 m be limited to 100 kW. Russia was excepted from this recommendation at her own request.

As regards stations using waves above 1 000 m, the liability of interference with air services was considered and certain articles of the Washington Convention were reiterated which forbid any increase of power of long-wave broadcasting stations if such increase should cause interference with existing services.

Some Problems of the Immediate Future.

One of the most interesting problems is to determine what application can be made of the band of higher hertzian frequencies of the order of 30 000 kilocycles per sec. and above. At the present time they are mainly employed for such purposes as rotating beacons for ship navigation and for short-distance communication, but further developments of aerial and reflector transmitting systems, together with receiving apparatus, may make them suitable for highly directional communication systems.

In regard to the receiving problem on these waves, it is important to emphasize the need for valve development in which the inter-electrode path is reduced to a minimum; for the time of flight of the electron at the higher values of these frequencies, such as those of the order of 300 000 kilocycles per sec., is comparable with the period of oscillation.

The work of the C.C.I.R. to which I have referred brings out the necessity of aiming at the highest possible degree of frequency stabilization and reduction of permissible tolerances to admit of an increased number of services without interference. In this connection the importance of developing suitable monitoring stations and apparatus for the checking and setting of transmitters to definite frequency values cannot be overestimated. It is interesting to note that at the present time the technique of frequency control of short-wave transmissions approximates to the accuracy with which the absolute value of the frequency can be measured.

The further development of receiver control, for example the method of " gain control " used at the present time for counteracting the effects of fading, is an urgent requirement for short-wave services. The development of the screened-grid types of valves having better effective screening might lead to considerable improvement as regards amplification and to better performance results on the shorter band of wavelengths.

Turning to the atmospheric interference problem, as regards long-wave reception the improvements of receivers with a view to reducing this have been small, and this interference is still an outstanding difficulty. Advantage has, however, been obtained indirectly by the use of the higher frequencies of the short-wave band, where the atmospheric interference problem is not so serious.

The development of single-unit thermionic valves for the generation of very high powers is not proceeding so fast as was at one time to be anticipated, and considerable development work is still required. As regards the receiving valve, it would appear that there is still room for improvement in connection with the reduction of microphonic noises.

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NAVAL WIRELESS TELEGRAPH COMMUNICATIONS.*

By G. SHEARING, B.Sc., and Captain J. W. S. DORLING, R.N., Members.

(*Paper first received 5th July, and in final form 14th October, 1929; read before the* WIRELESS SECTION *4th December, and before the* HAMPSHIRE SUB-CENTRE *19th December, 1929.*)

SUMMARY.

The paper gives a short account of the development of wireless telegraph apparatus for naval purposes. The historical development is referred to, the wireless telegraph lines required for naval ships are outlined, and the essential requirements and chief features of some of the apparatus used under sea-going conditions are briefly described. The organization for giving effect to the policy of H.M. Board of Admiralty is also indicated.

Part (1). HISTORICAL.

The first experiments in wireless telegraphy in the Navy were commenced by Captain H. B. Jackson, R.N. (now Admiral of the Fleet Sir Henry Jackson) in H.M.S. " Defiance " in 1896. Trials were carried out in 1897 between " Defiance " and the gunboat H.M.S. " Scourge," when ranges up to 3½ miles were obtained. The transmitting apparatus consisted of a Hertzian oscillator energized by an induction coil worked either from the electric light mains through a resistance or from a 12-volt battery. The arrangement is shown in Fig. 1. A diagram of the receiving circuit is shown in Fig. 2, the detector being of the coherer type, a glass or ivory tube containing two metal plugs being employed. It will be seen from the diagram that arrangements were made for the automatic recording of signals.

As a result of these early experiments, attempts were made to develop greater power with a view to getting longer ranges, and in 1898 this development work was in hand, but trouble was experienced owing to the failure of the induction-coil insulation under damp conditions. This difficulty was eventually overcome by coating the coil with paraffin wax, and, in 1899, sets were fitted in H.M. Ships " Europa," " Juno " and " Alexandra," Mr. G. Marconi (now Marchese Marconi) being in charge of the trials, which were carried out during the peace manœuvres. Effective ranges of 50 miles were obtained during these trials. The collaboration between scientist and naval officer, started at this early stage persists to the present day with equally fruitful results.

In 1900 further progress was made and by the end of the year 43 ships and 8 shore stations either had been, or were being, fitted with wireless telegraph apparatus. Thirty-two of the sets employed were supplied by the Marconi Co., the remainder being constructed in H.M.S. " Vernon " and H.M.S. " Defiance " to Captain Jackson's specifications. The speed of signalling was about 10 words per minute and reception was automatic.

During 1903, Muirhead's automatic transmitting apparatus was installed for trial. Only two waves were in use at this time, these being known as " A " tune and " B " tune, " A " tune being of 395 ft. and " B " of 1 150 ft. A variable-wave set known as " C " tune was developed at a later date. The magnetic detector was tried and introduced at this period.

In 1905 a naval wireless telegraph establishment was organized in " Vernon " and work was commenced on an experimental 5-kW set for erection at Horsea Island in Portsmouth Harbour. In this year the well-known rejector circuit for reception was introduced by Mr. H. A. Madge, who had recently been appointed as civilian wireless expert in " Vernon."

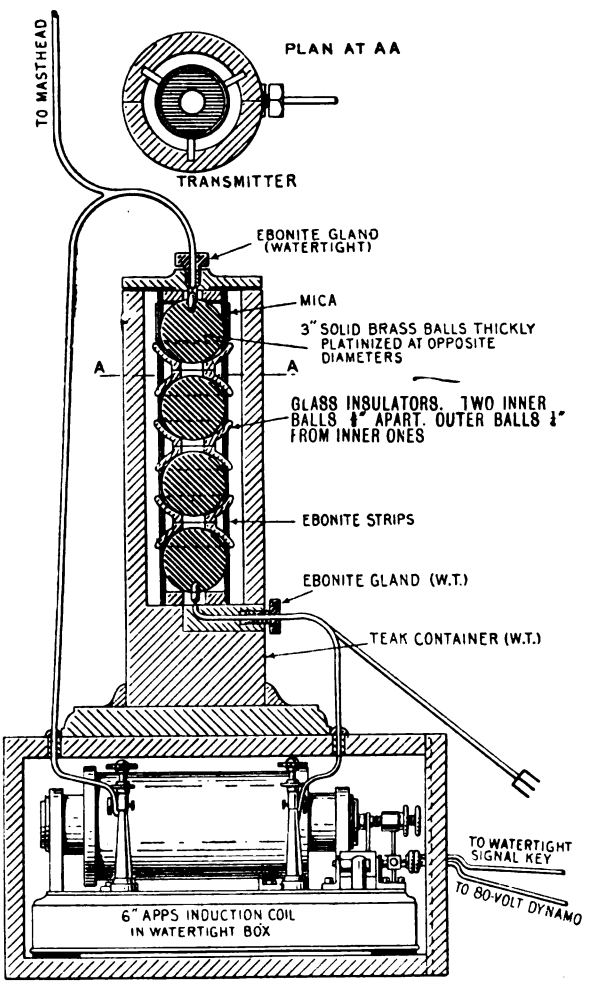

FIG. 1.

At the end of 1905, naval interest in wireless telegraphy was somewhat on the wane owing to the very large

* Reprinted from *Journal I.E.E.*, 1930, vol. 68, p. 237.

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number of failures which had occurred during trials at sea. At the beginning of 1906, however, a series of competitive exercises was carried out in the various fleets, with more satisfactory results. Up to this time, the development of wireless telegraphy had been under the control of the Torpedo Branch of the Navy, but signalmen had been used for operational purposes.

In 1906 it was decided to commence a new branch, the Wireless Telegraph Branch, in which men could specialize. This had very marked beneficial results, as would be expected since the candidates for the new branch were all compelled to undergo a special course of training in wireless telegraphy. Operators rapidly became more efficient and the number of failures during manœuvres consequently decreased.

During 1906, experiments were carried out on a number of new wavelengths. Towards the end of the

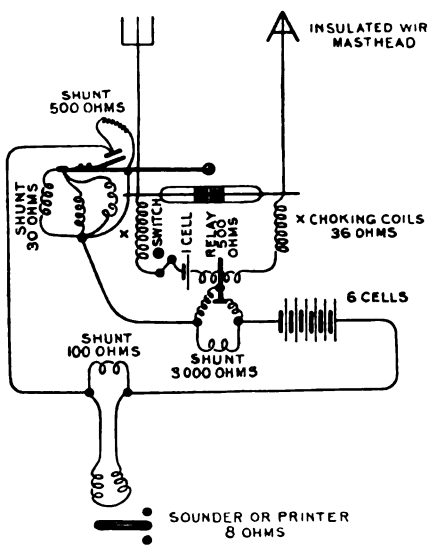

FIG. 2.

year, range trials were carried out between "Vernon" at Portsmouth and H.M.S. "Furious" on passage to the Mediterranean. Ranges of 600 miles were obtained regularly by day and up to 1 200 miles by night. The wavelength known as "R" tune, 2 500 ft., was found to be the most satisfactory. A long wave of 7 000 ft. was not found satisfactory. During this year, with more operators available, telephone reception became the normal procedure and the silent cabinet for reception was introduced in ships.

In 1907 the service Mark II set, later known as Type 1 spark, of the order of 12-kW H.T. output, was introduced for heavy ships, and, in 1908, experiments were commenced with the Poulsen arc in "Vernon" and "Furious." Preliminary experiments with wireless telephony and various forms of crystal and electrolytic detectors were also commenced in this year.

In 1909 the crystal detector was first used. High-power shore sets employing spark telegraphy were put into operation at Horsea, Cleethorpes and Gibraltar, and medium-power sets at Pembroke, Aberdeen and Ipswich. In 1910, a quenched-spark set was built for the naval airship. In this year also a wireless telegraphy course for officers qualifying for the Signal rank of Lieutenant (S) was approved.

During the years 1911–1912 a very considerable amount of experimental work was done, using the Poulsen arc, but there was no standardization of the apparatus at this period. The question of the development of methods of accurately measuring wavelengths also received considerable attention. During these years, also, the development of spark telegraphy sets continued and by 1913 the following were the main ship sets then in use:—

Type 1 (old Mark II main-power set).
Type 2 (old Mark I original low-power set).
Type 3, battleship auxiliary set.
Type 4, destroyer set.
Types 5 and 6, portable and harbour defence sets.
Type 9, cruiser auxiliary set.
Type 10, submarine sets.

Of these sets, Type 1 utilized a d.c. power supply of 20 kW, the remainder 3 kW or less.

During 1913 a 100-kW Poulsen arc was installed at Horsea Island, and three 440-ft. wooden lattice masts were erected, this work being carried out by Mr. C. F. Elwell. Also during this year the use of heterodyne reception for continuous waves was commenced.

During 1914, immediately prior to the commencement of the War, experiments were in hand on valve reception and arrangements were being made for complete arc sets to be supplied to ships at sea. The commencement of the War gave a great impetus to the development of wireless telegraphy, and the development of the thermionic valve is mainly associated with the war period from 1914 to 1919, first as a method of reception and later as a transmitter. In 1917 the wireless telegraphy established in "Vernon" was transferred ashore to H.M. Signal School, Portsmouth.

Whilst spark and arc systems of wireless telegraphy were mainly employed by ships at sea during the war period, the experimental development of the valve during the same time indicated that there was a probability that both these systems would ultimately be superseded by the thermionic valve system. In January, 1919, aerial currents of 50 amperes were obtained using a shore station valve transmitter.

The post-war developments which have occurred since 1919 have confirmed this view, and at the present time, except for special emergency purposes, spark transmitters are becoming obsolete and arc transmitters have been almost entirely replaced by valve transmitters in all types of ships. At the conclusion of the War, sets employing valves for power transmissions utilized valves of the order of 400 watts anode rating, but since this period the development of power valves has been rapid, until now single valves are in use with sets at sea which are capable of dissipating 8–10 kW at their anodes under oscillatory conditions.

The high-power sets developed up to 1923 were long-wave transmitters for long-range working, but during this year long-range experiments at sea in connection with the higher frequencies above 3 000 kilocycles

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per sec. (100 m) were commenced. During cruises in 1924 and 1925 of H.M.S. "Yarmouth" to the Canary Isles and to the Mediterranean, it was found possible to maintain short-wave communication with ships at Portsmouth during certain periods of the day, and in 1925 these experiments were continued with good results at greater distances to the Far East. The world cruise of H.M.S. "Renown" during January to June, 1927, provided another opportunity for testing the capabilities of short-wave working on long ranges, and except for three days in a period of six months it was found possible to maintain communication during

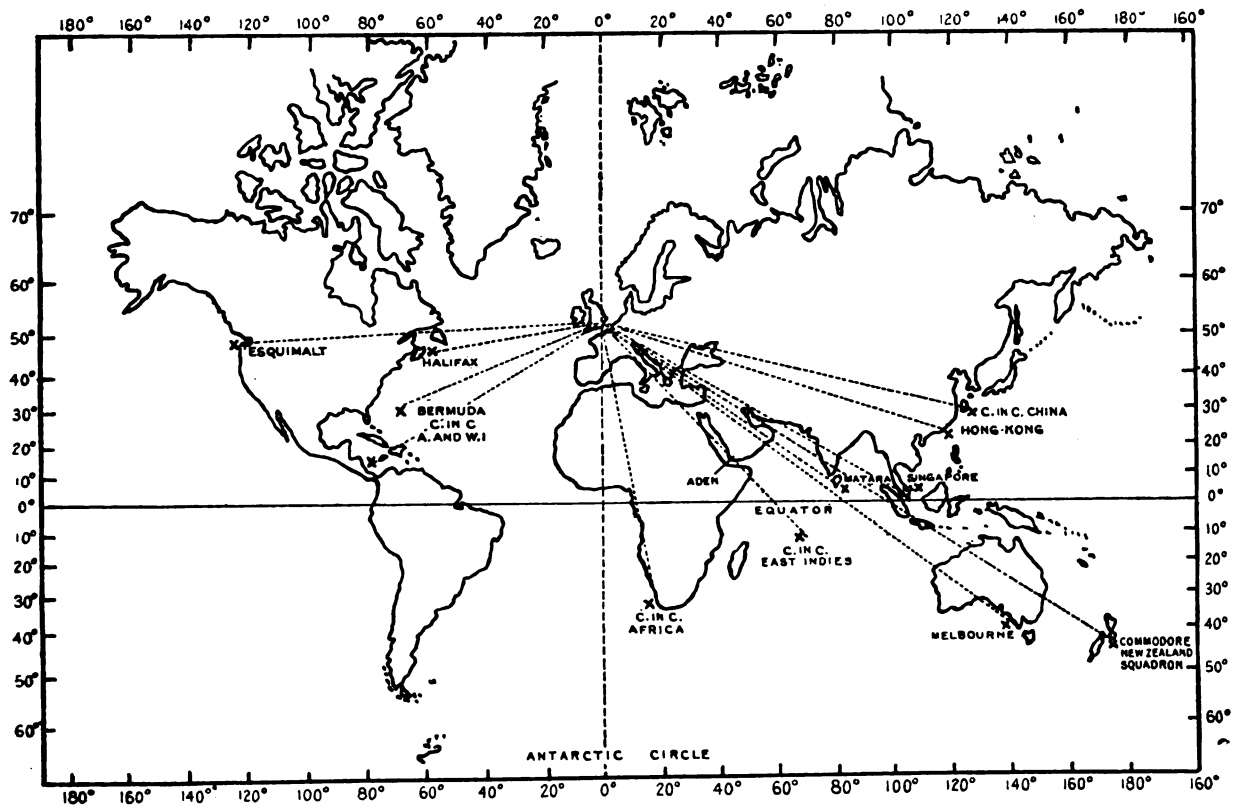

FIG. 3.

a part of each day of the cruise between Whitehall and "Renown," Whitehall controlling a short-wave transmitter at Horsea Island.

Regular service routines employing short waves between England and naval ships and shore stations overseas have now been in operation since 1926.

Apart from the direct application of wireless telegraphy to communications, the apparatus developed by H.M. Signal School for distant control, by the use of wireless signals, of target ships, such as H.M.S. "Agamemnon" and H.M.S. "Centurion," is well known and has been referred to in the Press.

Part (2). REQUIREMENTS TO BE FILLED BY A NAVAL WIRELESS TELEGRAPH SYSTEM.

It is a truism that trade follows the flag. The reverse that wherever trade flows there must the flag be also, is equally true.

The trade of this country is world-wide, and hence British men-of-war are to be found in all parts of the world. These ships are divided up into squadrons and are normally stationed in certain areas, for example China, East Indies, Africa, America and West Indies. Each squadron is under the orders of its Naval Commander-in-Chief.

A system of communication, to be effective, must link up the ships of these far-flung squadrons with their own Commander-in-Chief and with the Admiralty. The backbone of the system consists of a series of naval shore wireless stations situated in various British possessions throughout the world and in most cases forming a continuous chain of communication as indicated in Fig. 3 (Horsea and Mediterranean stations are not shown).

NAVAL SHORE WIRELESS STATIONS.

In the United Kingdom there are therefore wireless stations designed to communicate with ships in adjacent waters, and others for really long-distance communication with the stations abroad. Each naval area or station abroad has, depending on its size and position, one or more wireless stations, which act as terminal points for the traffic to and from the Admiralty and other commands, and as distributing and collecting centres for ships in the vicinity. Where commercial wireless stations are not available the naval stations are used also for communicating with merchantmen.

The wireless telegraph apparatus of these shore

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stations has therefore to be capable of communication with:—

(*a*) The Admiralty, either direct or through links.
(*b*) The Commander-in-Chief of the station and the ships on his station.
(*c*) Any merchant ships in the area.
(*d*) Adjacent naval commands.

To fulfil these requirements a medium-power long-wave valve transmitter of about 25-kW H.T. input is usually fitted for local working within a radius of, say, 1 500 to 2 000 miles. For the longer distances short-wave transmitters of various types are used. These range from 2- to 10-kW H.T. input. Recently it has been the practice to put all the receiving apparatus in a separate building at some distance from the transmitters.

Arrangements are made whenever possible for the wireless stations to be linked up with the cable and land-line system.

Cruisers and Other Ships on Foreign Stations.

These ships must be able to communicate with:—

(*a*) The Commander-in-Chief of their station and other ships on the station.
(*b*) The shore wireless station in the area.
(*c*) Adjacent commands.
(*d*) Merchant ships and aircraft in their vicinity.

The advent of the short-wave system has made it possible for ships in any part of the world to communicate direct with each other and with the Admiralty at certain times of the day. This facility is made use of to supplement the normal lines of communication and to avoid delays to especially important messages.

To fulfil the requirements stated above, the wireless telegraph equipment in H.M. ships of cruiser class and above has the following characteristics:—

(1) Range (long wave)—1 000–2 000 miles.
(2) Power (long wave)—12-kW H.T. input to transmitter.
(3) Power (short wave)—8-kW H.T. input.
(4) Valve transmitters are used in all cases.
(5) Receiving apparatus (long- and short-wave). Two or three sets are fitted.
(6) Direction-finding apparatus on the Bellini-Tosi system is fitted in most cruisers.
(7) Fire-control wireless sets are fitted for use when cruisers are concentrated in squadrons.

Apart from its actual technical performance, the apparatus must have certain very definite characteristics to fit it for use in a ship at sea. For example, it must be robust, simple to handle, reliable under all climatic conditions, proof against vibration and the shocks of gunfire, and capable of being maintained in full efficiency for months at a time by the ship's wireless staff.

The Main Fleet.

So far we have only considered the shore wireless stations and the cruising ships on foreign stations. Behind the cruisers and the ultimate source from which they obtain their liberty of action is the main battle fleet—the Grand Fleet of the Great War.

A self-contained battle fleet consists of battleships and battle cruisers which provide the main fighting power, together with cruisers for scouting, aircraft carriers, destroyers and submarines.

The whole force is under the command of its Commander-in-Chief, who must be able to transmit his orders to any unit of the force. The scouting units of the fleet must be able to transmit the information they obtain to the Commander-in-Chief and the other senior officers of the fleet. The Commander-in-Chief must also be in touch with the Admiralty and other authorities.

All these lines of communication should be able to work simultaneously without interference. This is usually done by multiple reception in the flagship, all transmissions being made by wireless guardships specially allocated for each wave. The apparatus used for these purposes is similar to that fitted in cruisers, already outlined.

Short-distance apparatus is also fitted in special offices in the battleships and cruisers to permit of intercommunication within the fleet without interfering with the long-distance lines. This short-distance apparatus is used for manœuvring the battle fleet and also provides an essential link between the wireless guardships and the flagship.

The striking feature of a fleet action is the speed at which everything happens. The two fleets may be approaching one another at 40 to 50 miles an hour, and even with aircraft reconnaissance the time between first sighting the enemy and joining action may be only an hour or two. During this limited time the scouts of the fleet have to observe that part of the enemy fleet which may be in sight, determine their position, composition, course and speed, and then signal this information to the Commander-in-Chief. From all these reports he has to determine what the enemy as a whole are doing and then manœuvre his force into the best position. The volume of signalling during this critical hour or so is very great, especially in the fleet flagship, and special means are necessary to ensure that no signal is delayed. The wireless apparatus is designed for the rapid handling of messages and so that the guardships can readily control the traffic on their particular wavelengths. In all ships the messages are received and transmitted, not in the wireless offices between decks, but in a remote-control office near the Admiral's or Captain's bridge, thus reducing to a minimum the intermediate stages between the issue of an order or report and its receipt by the person for whom it is intended. Pneumatic tube and buzzer systems are fitted for the rapid transmission of messages in case of a breakdown of the remote-control system.

The direction-finding sets fitted in some ships often prove their value during this stage of the action. Direction-finding bearings may be obtained of the enemy, or, should, for example, some of the scouting vessels or aircraft be uncertain of their positions relative to the battle fleet, a wireless bearing may be a useful check.

Wireless communication is also largely used after the action has started. Reports of the movements of the enemy forces continue to be made, and the fleet is manœuvred by a combination of wireless and flag signalling.

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All ships, down to destroyers, also use special low-power sets for fire control. The essential feature of this line of communication is that it should be free from interference from any other line and absolutely reliable even under action conditions.

For obvious reasons it is not possible to elaborate the above outline, but it will be sufficient to show that the requirements of a naval wireless system are many and varied, and that the overall efficiency of the Navy must depend largely on the efficiency of its wireless communications.

Part (3). OFFICES AND APPARATUS.

Typical Office Arrangements.—The initial lay-out of a modern warship involves an immense amount of preliminary work and trouble in the process of subdividing the available space for the various requirements. The fact that the ship is to be capable of fighting in action is a primary factor, and when provision has been made for the essential armament, propulsion machinery and living quarters for the crew, the space available for any other given requirement is very limited.

The ideal communication requirements are, of course, known, and a preliminary estimate of the space necessary is usually worked out in advance by the technical staff for the information of the Board of Admiralty. The actual space available is almost invariably found to be less than this estimate and it then becomes a matter of compromise as regards how far the ideal requirements may be met in the space which can be allocated. In this manner the wireless telegraph office is evolved of definite dimensions into which the main portion of the apparatus, apart from aerials, is to be placed. The general arrangements for the wireless telegraph offices and aerial systems of a battleship are indicated in the simplified diagram, Fig. 4.

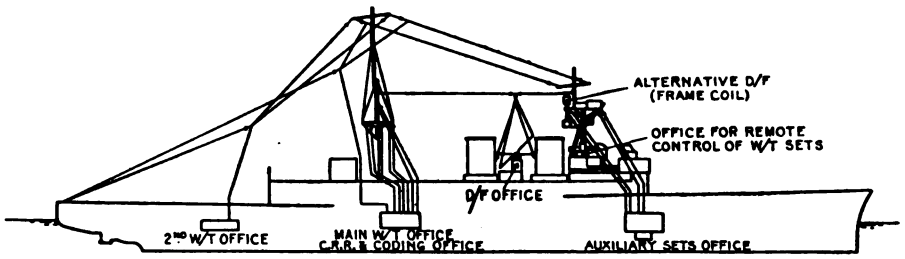
ALTERNATIVE D/F (FRAME COIL)
OFFICE FOR REMOTE CONTROL OF W/T SETS
D/F OFFICE
2ND W/T OFFICE
MAIN W/T OFFICE C.R.R. & CODING OFFICE
AUXILIARY SETS OFFICE

FIG. 4.

The main and auxiliary offices for heavy ships are situated below the main armoured deck, and the space has to include the following:—

(*a*) A silent cabinet for the accommodation of the receiving gear from which small trunks or cables go to the various receiving aerials.

(*b*) A cage protected by safety gate switches in which are contained the high-tension power supply and auxiliary apparatus for the transmitters. These have to supply oscillatory power to the aerial by means of a conductor fitted in a trunk which goes to the deck insulator to which the aerial system is connected.

(*c*) Low-tension power control apparatus for the wireless-telegraph motor-generators.

(*d*) Suitable gangways for ready access to all controls and for visual inspection of instruments.

(*e*) Coding office.

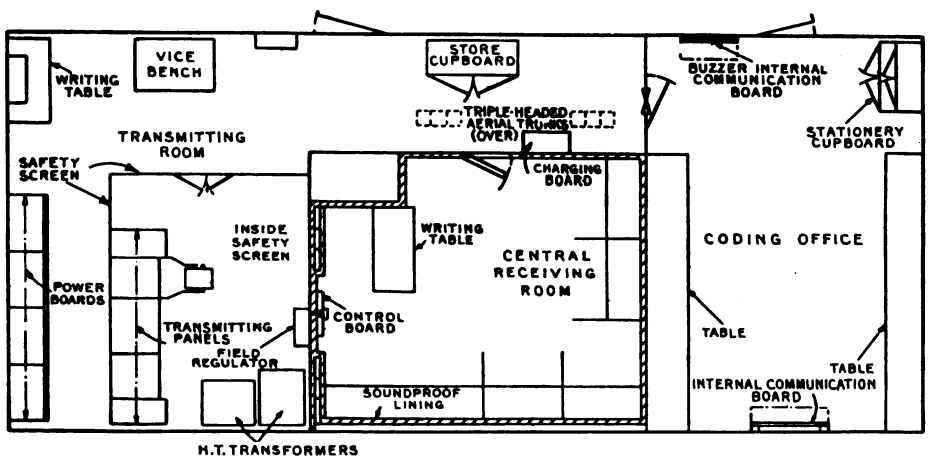

FIG. 5.—Typical lay-out of wireless telegraph offices in heavy ships.

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Primary Requirements.—The main conditions to which the wireless telegraph apparatus should conform are:—

(*a*) Space and weight limitations.
(*b*) Reliability of operation.
(*c*) Mechanical strength to resist shocks arising from gunfire and sea-going conditions.
(*d*) Ability to withstand extremes of temperature and weather.
(*e*) Efficient operation in offices below the main armoured deck, connected to the aerials via trunks or conductors of length which may exceed 100 ft.
(*f*) Minimum possible time of starting up for transmissions, with listening-through arrangements.
(*g*) Safety for operators. Controls readily accessible for quick operation.
(*h*) Rapid tuning arrangements covering a wide band of waves.
(*k*) Simple and easy to operate.

Design and Development.—The problem of designing oscillatory apparatus to fit into a given space which invariably is very much less than that available for the

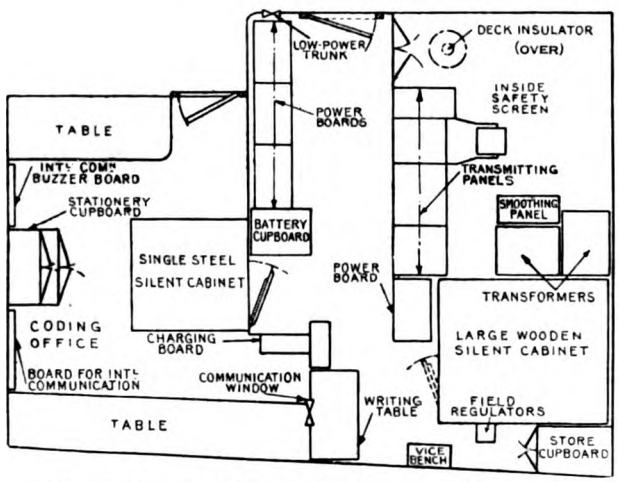

FIG. 6.—Lay-out of wireless telegraph office for light cruisers.

same power in shore stations imposes limitations on the technical design. The circuits and adjustments for wave-change have to be reduced to the minimum possible, the arrangements for replacement of gear have to be easy and rapid and the indicating instruments have to be capable of being viewed from widely varying positions.

The procedure usually adopted is to develop experimentally the apparatus first in a wooden model of the office space. From this the proposals and specifications for fitting the set in ships are prepared. Lay-outs for typical wireless telegraph offices of various sets for battleships, cruisers and destroyers are shown in the plans, Figs. 5, 6 and 7.

Installation and Trials.—The installation of the sets is carried out by the contractor or dockyard under the supervision of H.M. Signal School, Portsmouth, the completion of the wiring and final tuning of the high-frequency apparatus being carried out by the co-operation of the Signal School and the ship's staff. This installation work includes the following:—

(*a*) Fitting lined receiving rooms (silent cabinets).
(*b*) Fitting safety screen and all apparatus components.
(*c*) Fixing ventilation fans, air supply and exhaust trunks.
(*d*) Construction of aerial trunk, deck insulators and aerial conductor system.
(*e*) Fixing of all moving machinery, which is fitted outside the office.
(*f*) Wiring work as specified, covering low-tension lead-cased or bare high-tension wiring.

Upon the completion of the installation, the ship's staff has to carry out the following duties:—

(i) Test all the running machinery.
(ii) Complete the high-tension and high-frequency wiring for the transmitters and receivers not shown in the specification.

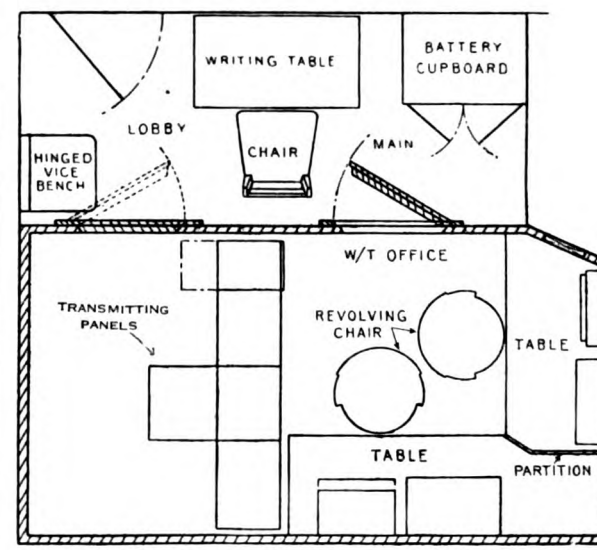

FIG. 7.—Typical lay-out of wireless telegraph office for destroyers.

(iii) Fill the oil-insulated apparatus such as transformers and condensers with oil which has been carefully dried.
(iv) Erect the aerial system.

The sets are then ready for tuning and external communication trials, which are carried out by Signal School Officers in conjunction with the ship's staff.

APPARATUS.

(A.) *Power Supplies for Main Wireless Telegraph Transmitters.*—The power supply for both thermionic valve or spark wireless telegraph transmitters is mainly high-tension low-frequency alternating current, and the earlier post-war development of valve transmitting sets was facilitated by the fact that existing power plant which was installed initially for spark transmitters could be used also for valve transmitters. The frequency value for the alternators has been determined mainly by

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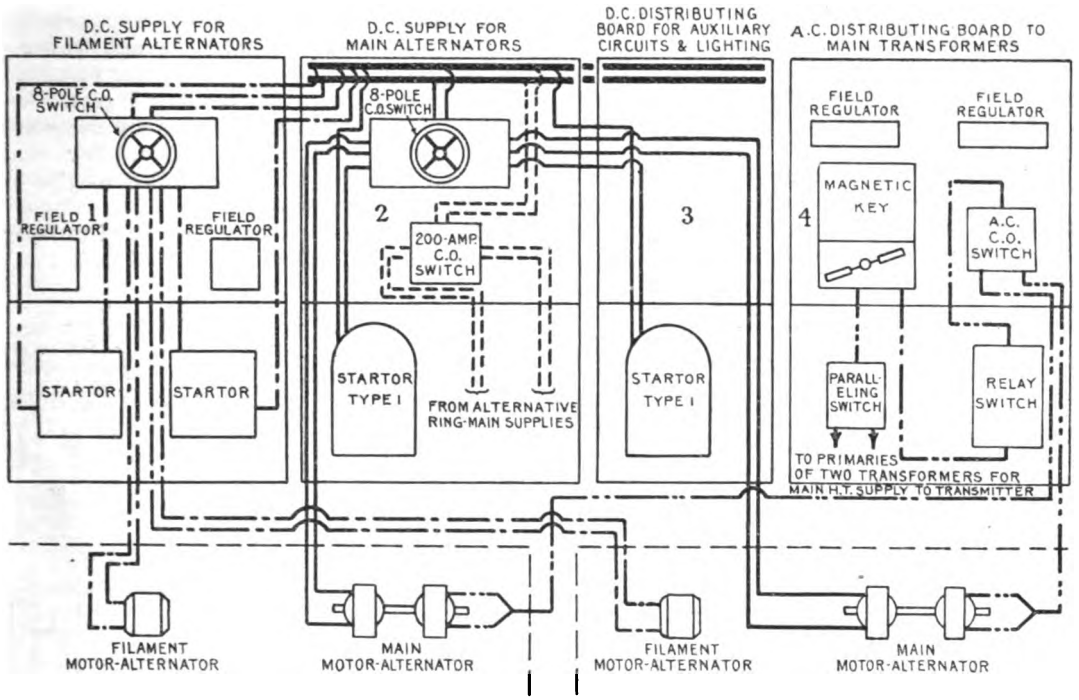

FIG 8.—Diagram of power boards and power arrangements.

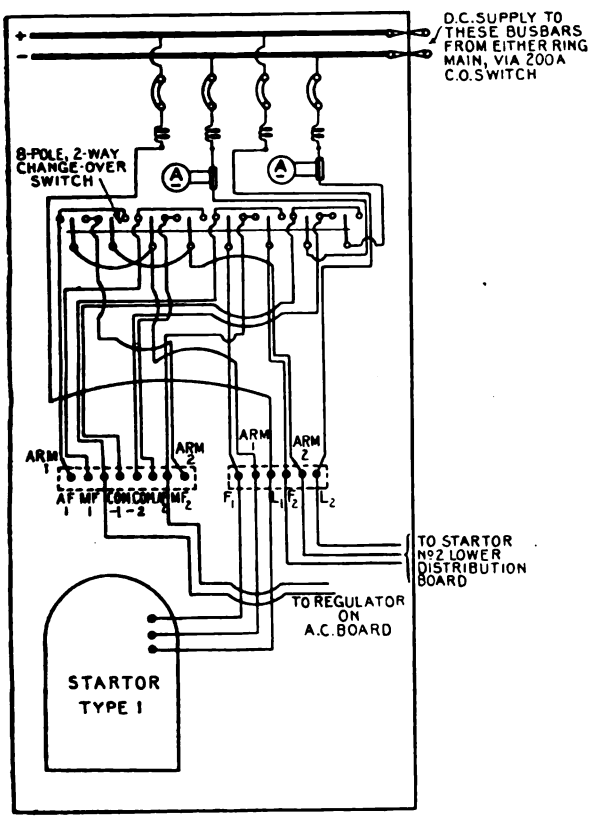
D.C. SUPPLY TO THESE BUSBARS FROM EITHER RING MAIN, VIA 200A C.O. SWITCH
8-POLE, 2-WAY CHANGE-OVER SWITCH
ARM 1
ARM 2
ARM 1
ARM 2
TO STARTOR No2 LOWER DISTRIBUTION BOARD
TO REGULATOR ON A.C. BOARD
STARTOR TYPE I

FIG. 9.—D.C. supply boards for main motor-alternators. Wiring diagram showing supply to main alternators.

Note.—Com. —1, Com. —2 denote common negatives 1 and 2.

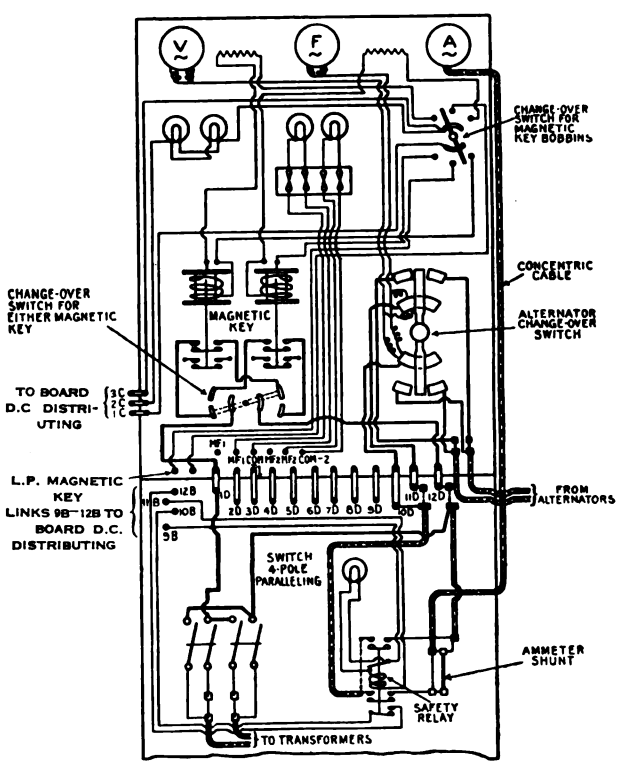

FIG. 10.—A.C. supply boards for main motor-alternator.

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the note frequency which can be read easily through other wireless telegraph interference and is not tiring to the operator when listened to for long, continued periods. For spark transmitters the frequency adopted was 350 cycles per sec., giving a note frequency of 700, but for the valve transmitters this has been changed to 500 and 1 000 cycles respectively.

The power supply for the motor-generators is taken from d.c. busbars on power boards in the wireless office supplied from the ring mains of the ship. This supply is at either 220 or 100 volts, depending on the ship.

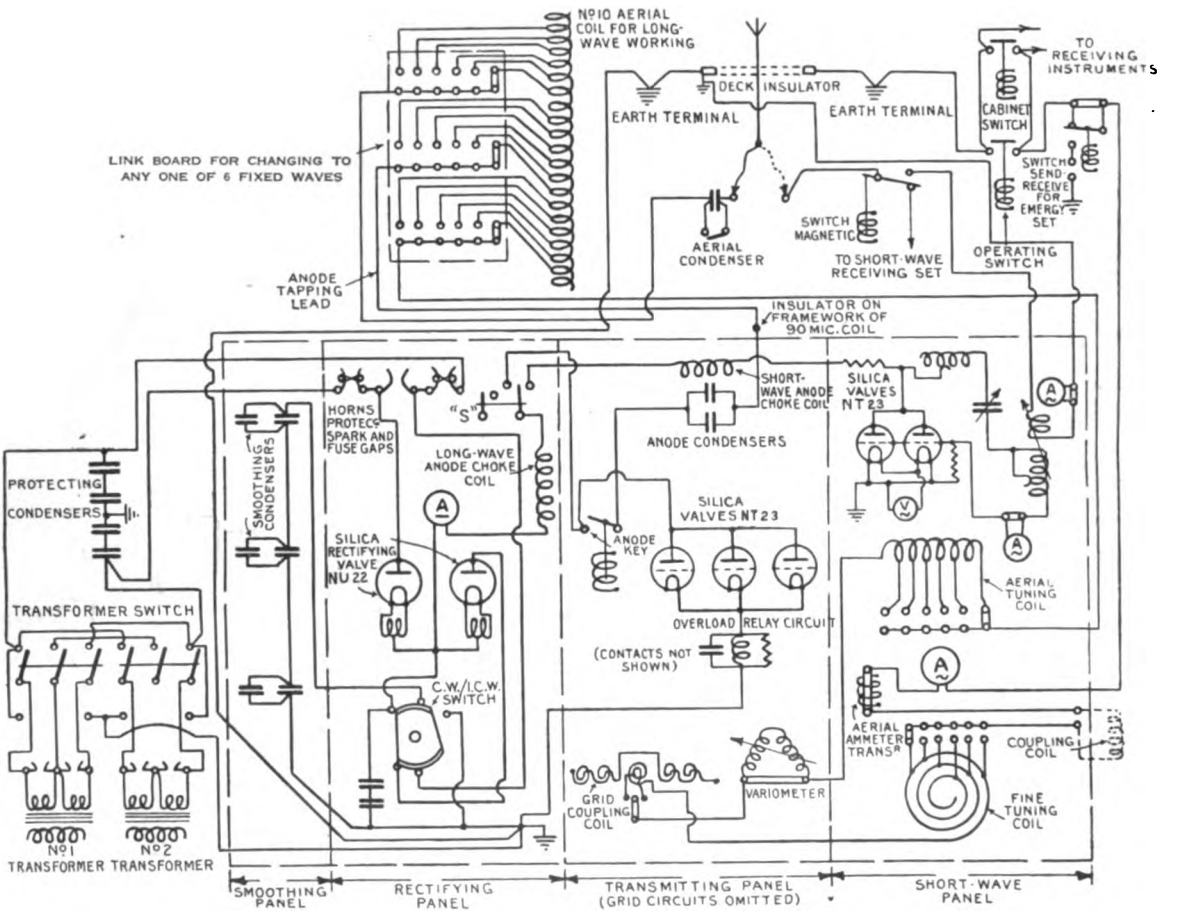

FIG. 11.—Diagram of H.T. and oscillatory circuits. "S" = change-over switch for long- or short-wave transmitter.
(Low-power panel not shown.)

The ring-main power supply is in duplicate and these supplies are via branch circuit-breakers fitted in separate watertight compartments outside the wireless telegraph office which are controlled by "on" and "off" pushes from inside the office. The machines are usually installed below the water line of the ship, with their axes arranged fore and aft. It is usual to mount one pair of machines, e.g. main and filament motor-alternators, in one watertight compartment separate from the other pair. The earlier machines were of three-bearing type, but these have now been superseded by four-bearing machines, the motor and alternator being separate machines with flanged coupling, due to overheating of bearings with the earlier type and the greater ease of passing the four-bearing machines, when separated, through the hatchways.

Duplication of the machines and power panels is adopted, each motor-generator being installed complete with its own d.c. and a.c. control gear. In addition, arrangements are also made whereby either machine can be operated from either startor. For certain cases automatic control gear is installed to give a constant value of generated voltage.

The power panels themselves are of metal framework structure and follow general electrical engineering practice for marine purposes. Each of the panels is contained in an angle-iron frame so arranged that any of the individual units can be readily removed.

Subdivision is carried to a greater degree than otherwise would be the case, to enable each part to pass through the hatches.

The power supply is so arranged that both machines may be running simultaneously, but if one is on load then the other must be running light. The power panels are comprised in four main parts which subdivide into two parts. These parts serve the following functions:—

(i) D.C. supply and a.c. output board for the filament motor-alternators.

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(ii) D.C. supply board for main alternators, with busbars for supplying No. 1 board.
(iii) Distribution board for lighting and all auxiliary circuits, including operating and control circuits, fans and motors charging battery circuits.
(iv) A.C. distribution board for main alternator.

A schematic diagram showing the general arrangements and functions of these parts is shown in Fig. 8.

The earlier hand-starting has been replaced by auto-

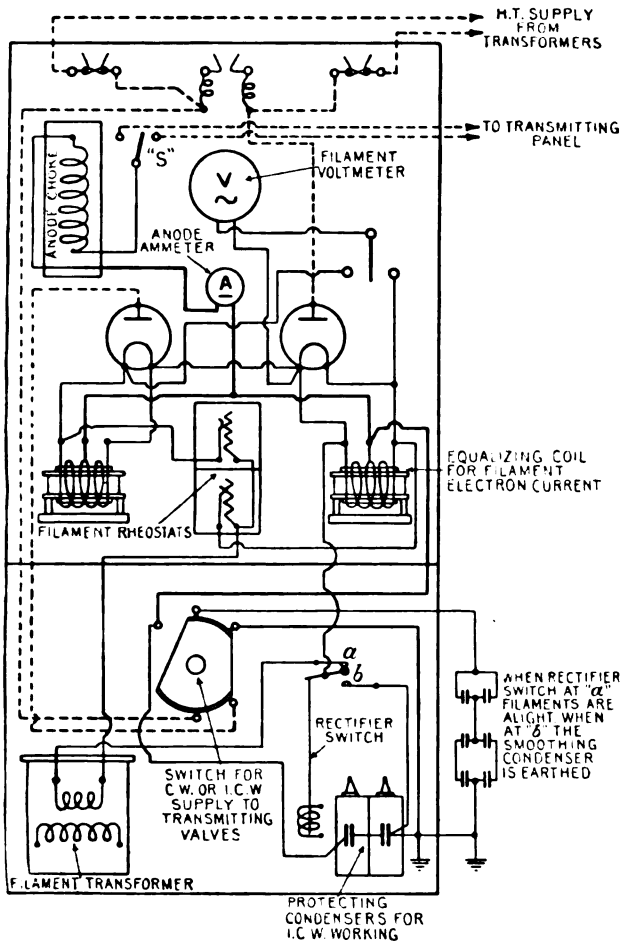

FIG. 12.—Rectifying panel.

startor controls and these have proved much more satisfactory than the hand startor. Tests have shown that the auto-startors fitted do not throw off under the normal vibrations of the ship or shock from gunfire. As typical circuit illustrations for the power boards, wiring diagrams for the d.c. and a.c. supply boards for the main motor-alternator are shown in Figs. 9 and 10. With regard to the d.c. board, duplicate supplies are taken from the busbars through pairs of circuit breakers. The positive supply leads go from each circuit breaker to one of the two startors for the motor-alternators. The negative lead goes to an 8-pole, 2-way change-over switch which serves the dual function of d.c. change-over for alternators and for startors, enabling either startor to be used with either motor.

The a.c. board supplies the primary of the main h.t. transformer. This supply is via a safety relay switch and magnetic key. Two magnetic keys are fitted and switching arrangements provided whereby either key can be brought into use for signalling with either alternator. The magnetic key is operated for signalling by solenoid control from the morse key.

The distribution boards also include indicating lamps and safety devices as follows:—

(*a*) Indicating lamps for main and filament d.c. motors.
(*b*) Indicating lamps for main alternators.

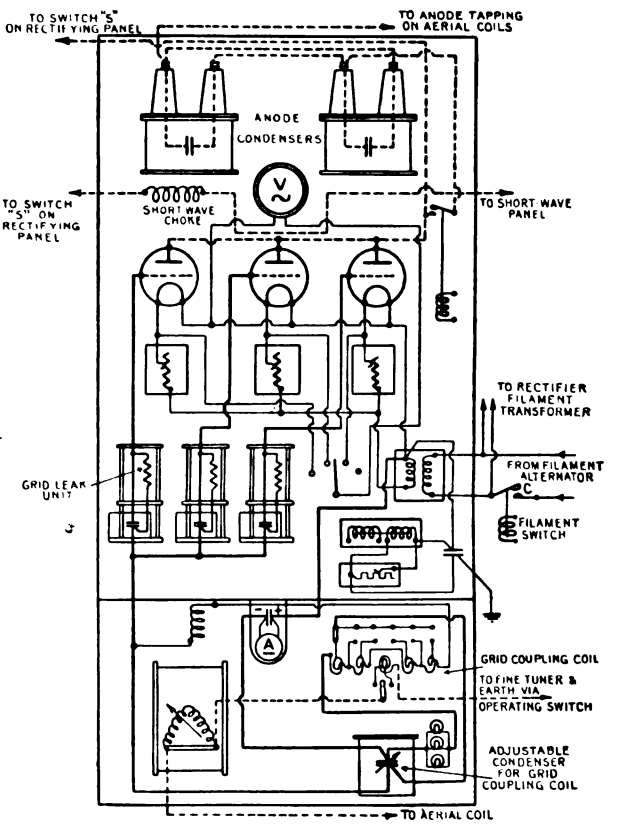
TO SWITCH "S" ON RECTIFYING PANEL
TO ANODE TAPPING ON AERIAL COILS
ANODE CONDENSERS
TO SWITCH "S" ON RECTIFYING PANEL
SHORT WAVE CHOKE
TO SHORT WAVE PANEL
TO RECTIFIER FILAMENT TRANSFORMER
GRID LEAK UNIT
FROM FILAMENT ALTERNATOR
FILAMENT SWITCH
GRID COUPLING COIL
TO FINE TUNER & EARTH VIA OPERATING SWITCH
ADJUSTABLE CONDENSER FOR GRID COUPLING COIL
TO AERIAL COIL

FIG. 13.—Transmitting panel.

(*c*) Internal signalling lamps between receiving office and power-boards, with bell and push on the board.
(*d*) Operating lamps.
(*e*) Locking-device for ring-main change-over switch.

(B.) *Design of Valve Transmitters for Long and Short Waves.*—It is not intended here to discuss the types of wireless telegraph circuits which are employed. The earlier post-war valve transmitters * were either directly connected to the aerial circuits or coupled via an intermediate or primary oscillatory circuit, these transmitters

* See G. SHEARING: "Naval Wireless Telegraph Transmitters Employing Rectified Alternating Current," *Journal I.E.E.*, 1925, vol. 63, p. 309.

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being used on what are usually known as long or medium waves, i.e. frequencies between 60 and 3 000 kilocycles per sec. (5 000 and 100 m). The later designs employ master-oscillator types of generator circuits, the frequencies in certain cases (for the shorter waves) being stabilized by quartz crystal control. The introduction of short waves has modified the requirements very considerably. Their advantages for long-range working at distances from a few hundred miles up to world-wide ranges during certain portions of the 24 hours of each day are well known, and it has been necessary to arrange for the fitting of attachments to existing longer-wave sets capable of transmitting on a band of waves covering 4 300 to 21 500 kilocycles per sec. (70 to 14 m).

The general arrangement of the oscillatory circuits for one of the earlier main valve long-wave trans-

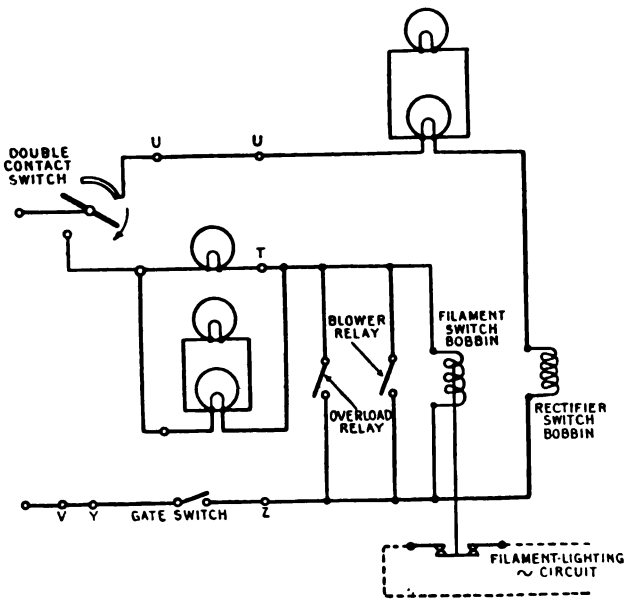

FIG. 14.—Simplified diagram of d.c. control and relay circuits for transmitting and rectifying valves.

mitters with short-wave attachment is indicated in Fig. 11. The rectifying and oscillatory circuits, with the exception of the aerial coils, are arranged in panels of light steel or aluminium alloy framework with the front and outer ends protected by expanded-metal screens. These panels are supplied with the internal wiring completed as far as practicable. The arrangement of typical rectifying and long-wave transmitting panels is shown in Figs. 12 and 13.

(C.) *Operating Controls for Valve Transmitters.*—For a valve transmitter as shown in Fig. 11, the following d.c. circuits are provided:—

(i) Signalling and remote-control circuits for operating all the switches working in unison with the morse key, e.g. magnetic key for main power to valve set, anode key for isolating the valve anode circuit from the aerial circuit, operating and cabinet switches which are closed for earthing the receivers during marking and opened for listening-through during spacing. The circuits of the remote-control bobbin and repeater to the bridge are also included. The morse key is fitted with a bobbin attachment for operation from the bridge and other remote positions.

(ii) Blowers for valve cooling and main safety circuits. The main power-supply relay is in series with

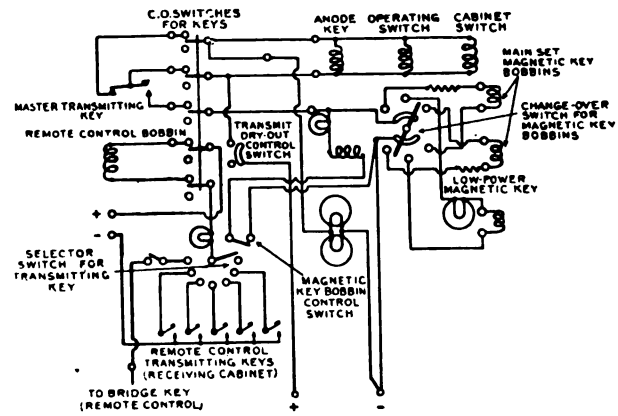

FIG. 15.—Diagram of signal and remote-control circuits.

the blower circuit when the gate of the cage is closed, so that it cannot be " made " until one of the blowers is switched on.

(iii) Alarm circuit. This warns the auxiliary offices when the main set is about to transmit. Indicating lamps are also fitted in the receiving cabinet, telling the operator of the main set when these auxiliary aerials have been earthed.

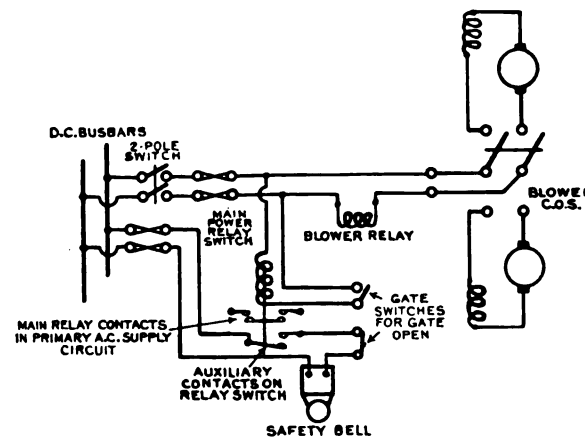

FIG. 16.—Diagram of blower and main safety circuits.

(iv) Internal signalling circuit. Lamps and bell with pushes for enabling the operator in the cabinet to convey his requirements to the operator at the power boards.

A simplified diagram of the valve filament and rectifier switch circuits is given in Fig. 14. The double-contact switch ensures that the rectifier switch contact " a " (see Fig. 12) makes before and breaks after the filament

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switch contact " c ", which is in the filament alternator supply lead to the primaries of all filament transformers (see Fig. 13). Thus no filament current is flowing when the rectifier switch is broken. If this were not done the contacts might fuse and prevent the switch from short-circuiting the smoothing condenser when the set is shut down after a transmission.

In addition, it will be seen in Fig. 14 that the contacts of the overload and blower relays are placed across the filament-switch magnetizing bobbin, so that unless they are open the filament switch is also open and the valves are switched off. The overload relay contacts close and switch off the valve filaments if for any reason whatever the valves carry excessive current. Similarly, the

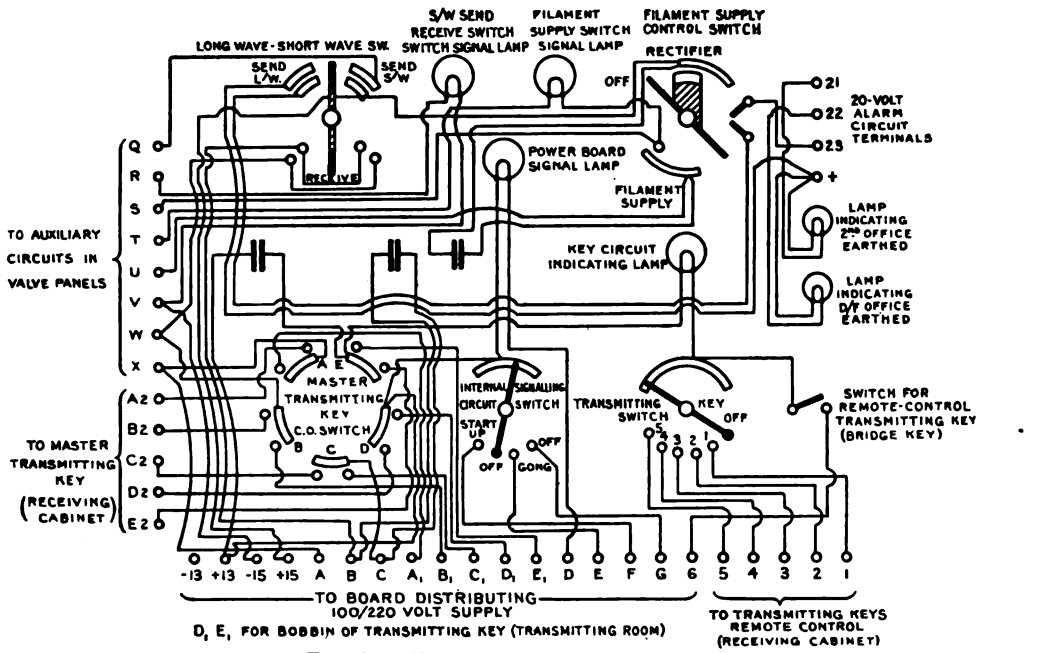

FIG. 17.—Transmitter control board.

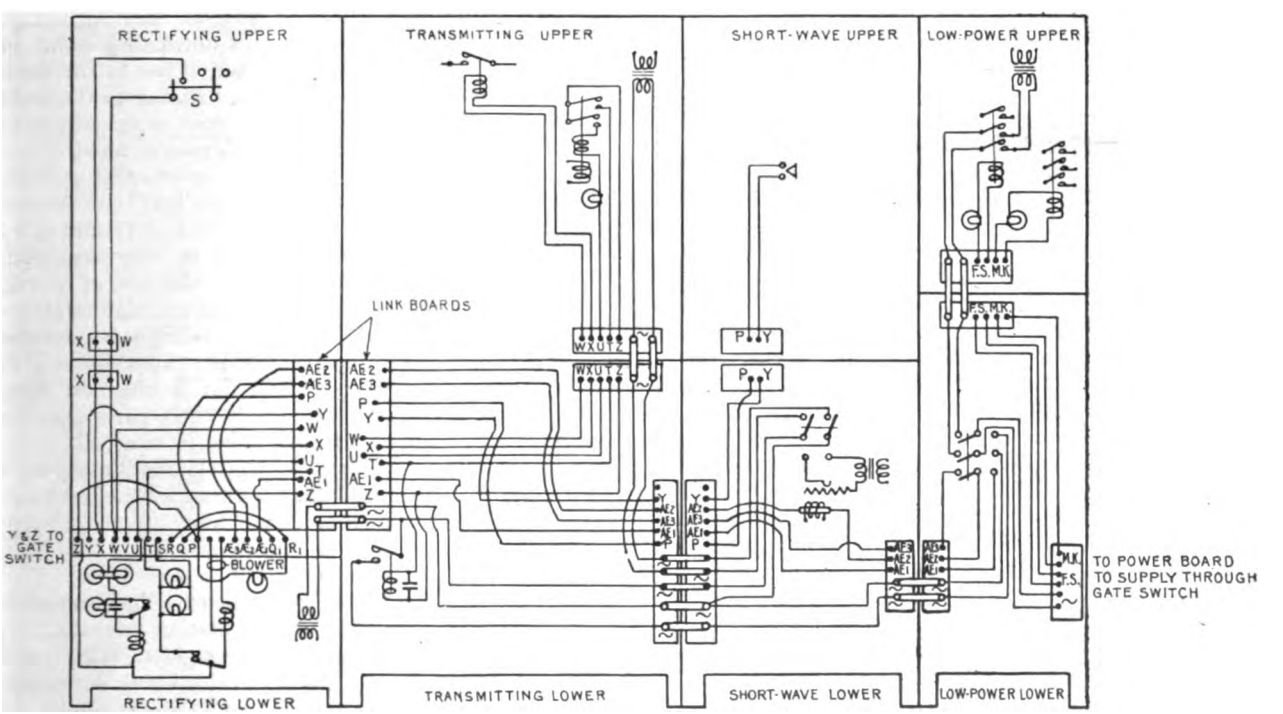

FIG. 18.—Auxiliary circuits in valve panels.

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blower relay-contacts close if the blower motor supplying cooling air to the valves is not running.

Signalling control of the main sets is usually provided for by alternative morse keys, one of which is installed in the central receiving room adjacent to the operator, and the other on the d.c. distribution panel adjacent to the set for testing purposes. A schematic diagram of this circuit is given in Fig. 15, in which are shown:—

(*a*) The keying circuits for operating the magnetic key used for signalling.
(*b*) Circuits for closing the anode key in the transmitter circuit, the operating switch in the aerial earth lead and the cabinet switches for earthing receivers during marking.
(*c*) The bobbin attachment for the morse key when the set is operated from the remote-control positions.

Fig. 16 illustrates the safety circuits, which are supplied from a 2-pole switch from the distribution panel. From this it will be seen that the main power relay is opened whenever the safety gates are opened. As a protection against signalling when a man is aloft, an indicator plate with the words " man aloft " is hinged to the power board in such a manner that when bolted down it fouls the switch handles of the 2-pole switch and so prevents it from being closed.

The alarm circuit for other wireless telegraph offices utilizes a separate d.c. 20-volt supply. In each office is provided a buzzer and lamp. With the closing of the main safety gates the buzzer sounds and the lamp glows. With the earthing of the office aerial the buzzer ceases to sound but the lamp continues to glow so long as the main set is in operation. A diagram of the operator's control board installed in the central receiving room is shown in Fig. 17. The method of arranging the panel d.c. circuits which are connected to the control board is shown in Fig. 18. Each separate panel can be transported with its control circuits completely wired, and links complete the circuits between panels.

(D.) *Valve Transmitters having High Constancy of Frequency.*—The more recent designs of combined long-wave and short-wave transmitters employ master oscillator control for both the long-wave and short-wave circuits. One of these circuit arrangements for a ship transmitter is shown in Fig. 19, in which the control stages are designed for operation on either long or short waves, and the main stage for short-wave operation comprises a high-frequency double-acting system of two valves which are neutralized to avoid retroaction. For long-wave working the valves are used in parallel. Signalling is obtained by grid-potential control. The general arrangements of auxiliary circuits are similar to those described in detail above for an earlier design. Switching arrangements are provided whereby the change from long-wave to short-wave working can be effected.

For shore station short-wave transmitters used for point-to-point working, quartz-crystal controlled transmitters are employed for frequencies of the order of 7 500 to 10 000 kilocycles per sec. (40 to 30 m). In certain cases, frequency-doubler circuits are employed for transmissions on the corresponding band of 15 000 to 20 000 kilocycles per sec. (20 to 15 m), so that one crystal controls transmissions on the two waves. A general diagrammatic circuit of such a transmitter is shown in Fig. 20, in which the amplifier circuits are of high-frequency double-acting design, the third harmonic of the crystal frequency being selected and amplified. Automatic signalling is provided for by grid-potential control of one of the lower power stages, using a Creed relay as indicated in the diagram. Fig. 21 shows circuit arrangements for doubling the frequency, i.e. to the sixth harmonic of the crystal frequency.

In the case of all naval ship transmitters, the circuit design is such that no spacing wave is transmitted and arrangements are provided for listening-through.

(E.) *Assembly of Transmitter Components.*—The main requirements to which the wireless telegraph apparatus should conform have already been enumerated on page 15, from which it will be seen that the assembly and design of the components are a compromise between electrical and mechanical requirements. It is especially important to ensure that the relatively weaker links in the apparatus must all be easy of access for removal for repair and replacement. In assembling the components the distribution of the electric and magnetic fields has to be borne in mind, and undesirable magnetic and electrostatic coupling between coils must be reduced to a minimum. Any long metal parts near high-voltage coils are so mounted that their axes lie as nearly as possible in an equipotential plane.

(F.) *Panel Frameworks.*—The shock of gunfire causes the decks and bulkheads of the wireless telegraph offices in some ships to whip by as much as ½ inch. A framework must therefore be bolted either to the deck or to a bulkhead, but not to both. Fixing to the deck is the best solution. Heavy parts such as transformers and oil-filled condensers are, for this reason, fixed to the framework base to minimize bending-moment stresses due to shock. The question of angle iron, non-ferrous metal or wooden frames for the panels is determined by the space available. If the space is very cramped, voltage considerations may compel the use of wood; if more space is available, then angle aluminium alloys have the advantage of lightness; also, being non-magnetic and of low ohmic resistance, they do not cause appreciable losses. Angle iron has strength but it must be lined with thin copper to reduce its losses where parts cross an intense high-frequency magnetic field from a coil.

(G.) *Transmitter Oscillatory Coils.*—The oscillatory coils used may be very briefly dealt with under the headings of coils used for (i) long-wave, and (ii) short-wave transmitters.

(i) *Coils for long-wave transmitters.*—The materials mainly employed for insulation are ebonite, paxolin, porcelain and pyrex glass. The trend of later design has been to build American whitewood frameworks to support combs of ebonite which carry the coil windings. Porcelain parts have also been used for the combs, these being made from plastic clay which is vitrified by firing, but they were found not to lend themselves to accurate manufacture as regards dimensions. Those made by the dust

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process have the disadvantage that they are non-homogeneous and are porous before glazing. Paxolin and similar products are only used for so satisfactory when subjected to the humid atmosphere in ships. The moisture absorbed by the surface of the wood increases its surface

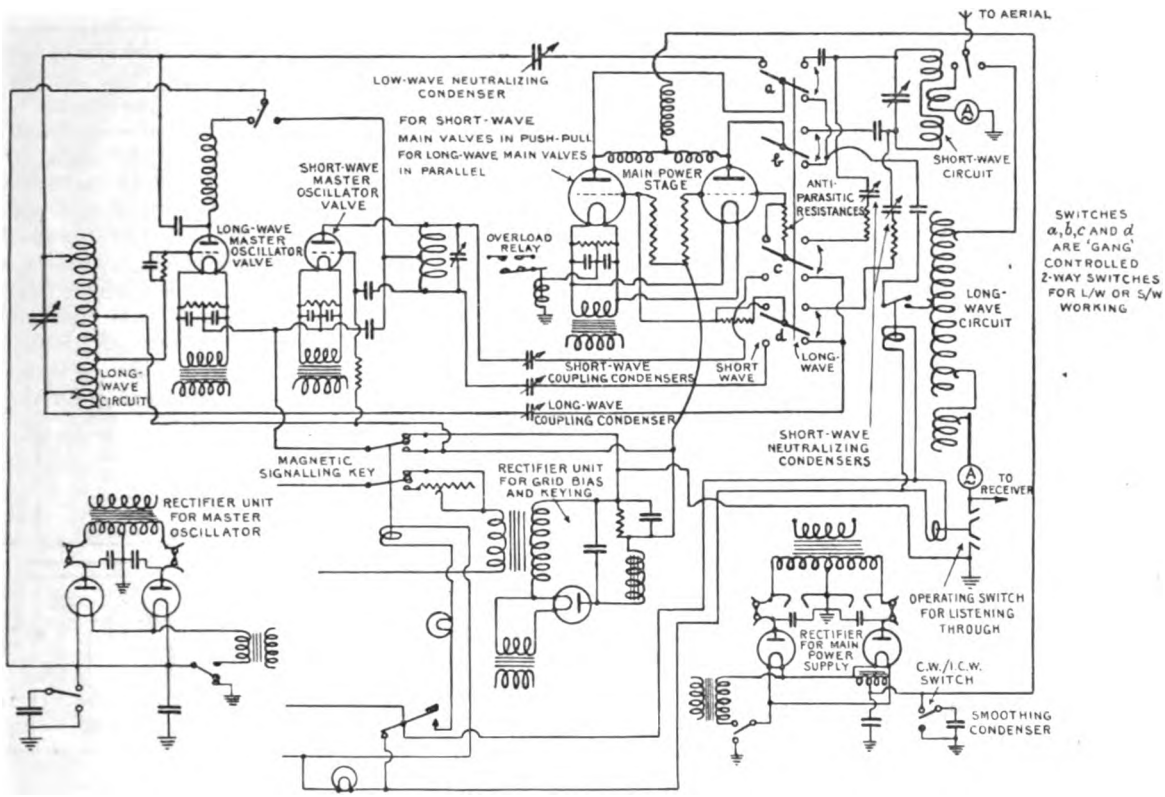

FIG. 19.—Naval type combined long-wave and short-wave master-controlled transmitter. Self-oscillating and low-power connections not shown.

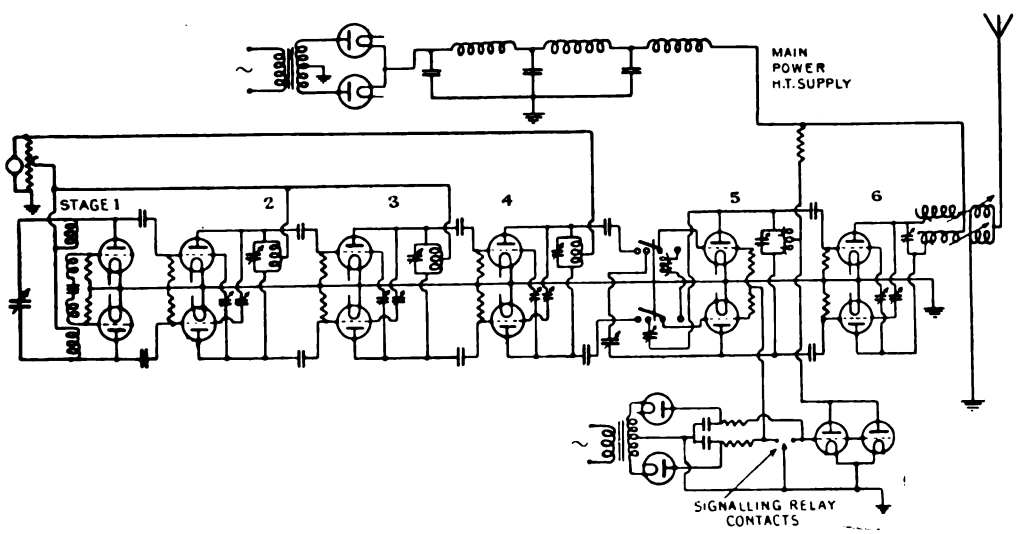

FIG. 20.—Simplified circuit diagram of quartz-crystal controlled short-wave transmitter.

parts requiring mechanical strength and not subjected to very high voltages at high frequencies. As regards American whitewood, whilst this is an excellent insulator when kept dry, it is not permittivity to such an extent as to cause brushing under the very high voltage gradients which exist for the coil designs due to space limitations. Treating the surface of the wood

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with a heavy coat of varnish largely destroys its high insulating properties, and the use of a very thin layer of varnish has been found to be not entirely satisfactory. As regards dimensions, a ratio of 3 : 1 in diameter to length gives the most efficient design of coil so far as copper losses and the length of wire for a given inductance are concerned. To obtain the required number of tapping points over a large range of frequency and to utilize the space within the framework to the best advantage, it is, however, usually necessary to reduce this ratio to 1: 1. In ship sets the long-wave coils are nearly always wound multi-layer because of space limitations. Stranded wire, each strand being insulated, is largely used and the windings are carried in slots. These slots are so designed that the voltage stress across a tooth does not exceed 5 000 volts per inch under the worst conditions. It is necessary to guard against the assumption that the greater the number of strands of the standard size (No. 38 S.W.G.) used in the wire, the less will be the resistance of the coil. For a given conductor there is a best spacing of the strands in the conductor and a best spacing of the conductors forming the coil. It is also desirable to keep the number of terminals down to a minimum. If a coil of N turns be tapped to form sections each having $\sqrt{N}$ turns, and if the last section is tapped at every turn, then the coil inductance can be tapped in steps of one turn throughout and the number of tapping points will be a minimum.

(ii) *Short-wave coils.*—For the larger ship sets these are wound with copper tube which carries a tapping clip for adjustment of wavelength. These coils are almost self-supporting but the end turns and a middle turn are usually clamped to porcelain insulators mounted on a wooden base, and in all designs the maximum permissible rigidity is necessary to secure constancy of wavelength. The electrical considerations tend to weaken the mechanical design, e.g. the turns have to be spaced sufficiently far apart to prevent the tapping clip from short-circuiting to the adjacent turns, and the coil has to be long compared with its diameter to enable single-turn tappings to be made having sufficiently fine adjustment of inductance in addition to the insulated clamping supports which are fitted to secure rigidity. The coils are also cooled by air blast to reduce the variation of frequency which otherwise occurs with change of temperature.

(H.) *Transmitting Condensers.*—The development of fixed condensers for specified wireless telegraph transmitting requirements is left to the makers of such apparatus. The usual type consists of copper-foil plates and ruby-mica dielectrics built up in sections connected in series or series-parallel as necessary to meet the specified conditions. Great care is taken to exclude any air bubbles between the plates and dielectric. The best procedure for assembly is to arrange for the sections to be placed in a vacuum in order to withdraw any air present within the condenser unit, and afterwards to dip them in paraffin wax for sealing. The procedure of placing the sections simply in boiling paraffin wax to drive out the air by heat is not satisfactory, for a proportion of the air may become sealed in spaces between the plates and dielectric at less than atmospheric pressure, and ionization losses may then occur.

The long-wave transmitting condensers usually have their units assembled in metal cases which are filled with either oil or wax and to which one of the condenser terminals is connected. The type of design which is favoured is that in which the case is suitably insulated from earth by porcelain pillars or blocks, by reason of the saving in space and cost for the container. It is of special importance to give the manufacturers full information as to the type of load and working conditions. The losses in a condenser are less when it is used in a spark or interrupted-continuous-wave circuit than for the case

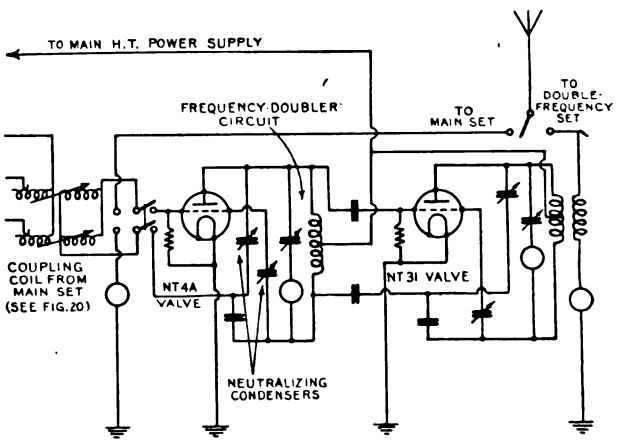

FIG. 21.—Frequency-doubling circuits for short-wave transmitter.

when the same circuit is used for continuous waves and passing the same effective current. The peak voltage values for the former are, however, much higher. The power factor of a good mica condenser is of the order of 0·2 of 1 per cent, but if the condenser is poorly constructed this power factor may easily increase to give double the power loss.

In connection with the losses in a condenser, the arrangement of the condenser plates is of great importance, for if the sections are so arranged that the bottom plates are connected to a low-potential terminal and the top plates to the high-tension terminal, the high potential gradient around the edges of the top plate may then cause heavy losses in the top clamping plate of insulation. Cases have been experienced where ebonite clamping plates have melted locally, adjacent to the edge of the top plate. Designs in which the condenser is arranged in two sets of units connected in parallel so that the h.t. plates face one another at the centre are much better in this respect; also, metal clamping plates may then be used. In certain cases fixed condensers are employed in short-wave transmitting circuits and it is then essential to keep the current path through the terminals, leads and condenser elements as short as possible, otherwise the inductance of this path may become serious.

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The adjustable tuning condensers used either are oil-filled for long-wave high-voltage transmitters or employ air dielectric for short-wave transmitters. The oil-filled condensers for small transmitting sets are usually more troublesome than the larger units, owing to the fact that any fibres in the oil are drawn between the plates by electrostatic attraction, with a tendency to bridge the plates. The condensers for short-wave transmitters are of low capacity, and to get a reasonable variation it is essential to keep their minimum capacity as small as possible. The chief difficulty experienced with the short-wave condensers is that a small roughness on one plate may start a brush discharge and cause arcing-over at a comparatively low voltage. In the case of transmitting sets employing neutralizing condensers, there is also the difficulty of maintaining the capacity-to-earth ratio of each set of plates constant to

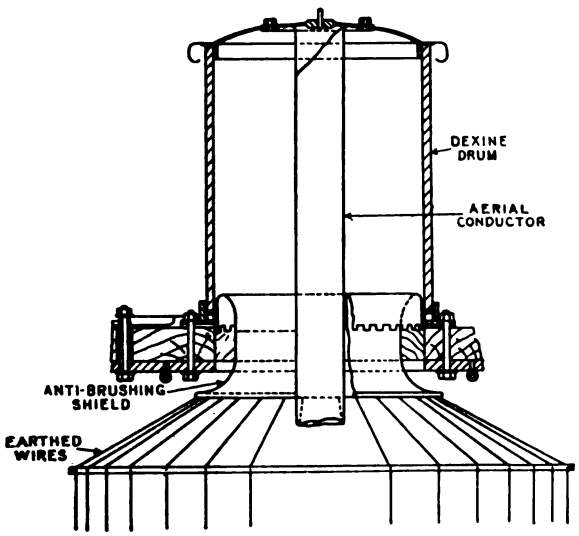

FIG. 22.—Deck insulator.

secure the symmetry which is essential for satisfactory operation throughout the tuning range.

(J.) *Main Aerial Trunk and Deck Insulators.*—The fact that the main wireless telegraph office is usually situated below the main armoured deck has necessitated the fitting of special trunks whereby the high-frequency energy may be delivered to the main aerial with minimum loss and without danger to personnel. The trunk itself is of steel, and to prevent loss of power due to eddy currents it was formerly usual to fit a cylindrical system of earth wires which were soldered to the copper lining of the deck insulators. These deck insulators were fitted on the upper deck and at the exit for the aerial conductor from the wireless telegraph office. In addition, a similar deck insulator was fitted at intermediate decks if these were armoured or formed part of the watertight subdivision of the ship. More recently the use of the earth wires has been abandoned. In some cases, however, the trunk is copper-lined. One design of deck insulator system is shown in Fig. 22.

The insulating material used in this particular arrangement is dexine, which has the disadvantage that if it becomes encrusted with salt, owing to exposure to spray, a burn-out is liable to occur. If, however, the burn-out is not too deep, so that the charred material can be removed by scraping, and the surfaces are then varnished with shellac, the insulator can be used again. More recently a porcelain bowl-shaped insulator has been used with suitable fittings for securing it to the lower deck casting as shown in Fig. 23. The upper casting and the central cylindrical conductor are at the aerial high-frequency potential, the copper lining being an anti-brushing shield of spun copper fitted through the hole in the deck. The aerial feeder inside the trunk is a copper tube of about 2 in. diameter. For the deck-insulator design of

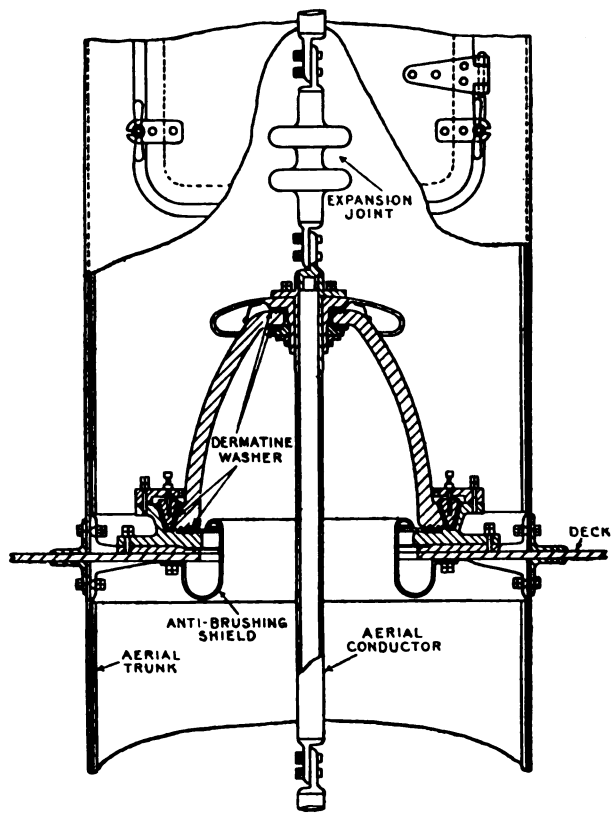

FIG. 23.—Deck insulator.

Fig. 22 the joints between the insulating drum and the castings are made by means of a bitumen compound to ensure the arrangement being watertight; dermatine rings are used for the design of Fig. 23. It is important to note that the insulator itself is not of sufficient strength to withstand a considerable pressure in an upright direction, and for this reason the steel trunk itself is designed to be a watertight structure with all doors and manholes into it fitted as watertight doors.

(K.) *Main Aerial Systems and Insulators.*—Two types of aerials have been employed in main aerial systems, these being either:—

(i) Sausage aerials having two or more legs each comprising 10 wires on circular spreaders at about 1 ft. diameter, or

(ii) Flat roof aerials of about 6–8 single wires in parallel, arranged along yards of 30 to 40 ft. in length.

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Experience with these two types indicates that for a given range and capacity there is very little difference in the communication, and the mechanical considerations of handling, rigging and maintaining the aerial in position are the main deciding factors as regards which type of aerial is employed. The chief advantages of the flat roof over the circular-spreader type of aerial are the reduction of strain on the masts for a given aerial sag and windage, the greater ease of handling and the reduction in cost. The wooden yards used for supporting the aerials are painted with an aluminium paint, thereby forming a metal sheath over the wood and preventing charring around the points of metal bolts or screws, which may otherwise be caused by brush discharges from the high-frequency field.

The run of the aerial feeders is of very great importance for efficient working. They must be set well clear of all obstructions and, especially, kept as far as possible from the tops of the funnels to avoid (*a*) reduction of effective height, and (*b*) the losses which otherwise may occur due to the low resistance from ionization of the gases. The upper deck insulator is provided with a spray shield formed by a light sheet-steel screen of about 5 ft. diameter around the deck insulator, and a special outhaul is necessary to ensure that the feeder is well clear of the top edge of the spray shield, unless the adjacent superstructure can provide a suitable attachment. The design of aerial systems for submarines, although they are smaller than those for main aerials of heavy ships, is in some respects more complicated, due to the fact that provision has to be made for raising and lowering the main aerial, which is only employed for surface working, so that it may be taken down and housed prior to diving. The mast is usually of telescopic design. For instantaneous working immediately after surfacing from submerging, a smaller (fixed) aerial is employed.

The strain and rigging insulators at present in use are of porcelain. This material, if properly glazed, has good electrical qualities for high-frequency working, and one of its chief advantages is the ability to withstand sea weather conditions and arcing over its surface for long periods without injury. Its mechanical strength is not so satisfactory. It is especially liable to fracture under mechanical shock, and cases are known where aerial systems have collapsed for this reason. In some cases where the porcelain insulator is cemented by the use of solder into metal end-sockets, failure has occurred due to melting of the cement from the heating of funnel gases, and special designs have been prepared to avoid this danger. In a few cases, also, a curious defect has occurred in some porcelain insulators. If the temperature is a little too high when stoving the glaze during manufacture, very minute bubbles appear in it. If some of these bubbles break and the salt from sea spray lodges in the resultant holes, a very serious reduction in the insulation resistance may be brought about. For this reason there is a tendency to use pyrex glass in place of porcelain. Tests of insulating materials are continuously in hand with a view to obtaining, if possible, a material having both good electrical and mechanical qualities.

The requirements of a good aerial strain insulator may be summarized as follows:—

Electrical.—High surface and volume resistivity. Low dielectric hysteresis loss. Preferably, low permittivity.

Mechanical.—Mechanically strong under tension. Must not be brittle. Surface must not deteriorate owing to action of funnel gases and/or immersion in sea water. Must be non-hygroscopic and the surface should be such that the settlement of moisture does not produce a continuous film. Surface must not be damaged by electrical flash-over.

(L.) *Receiver Power Supplies and Arrangements.*—The main receiving apparatus for heavy ships is installed in a silent room which forms a portion of the main wireless telegraph office. It has been laid down as a general policy that all receiving instruments shall be operated independently of the ship's mains, and the power supplies for the anodes and filaments have for this reason been taken from accumulators. Two complete sets of batteries are supplied for each office. The power for charging these batteries is taken from the 220- or 110-volt d.c. ship's mains. Motor-generators are installed for charging the low-voltage filament batteries, the anode batteries being charged directly from the mains through a suitable resistance. The control gear with fuses is mounted on a special switchboard which also contains the switches for the auxiliary 20-volt ship's supply for operating the buzzer circuits and supplying the secondary lighting circuit for the wireless telegraph office.

One common filament battery and one common anode battery are employed for all the models in a single receiving outfit. The earlier designs of models used bright-emitter valves employing 6 volts for the filament battery, but these models are being gradually converted to use a special dull-emitter valve having similar characteristics. The newer models have been designed to employ types of manufacturers' dull-emitter valves which have been found to be satisfactory, and a general review of the market led to the decision to adopt a filament battery of 4 volts. The common anode battery has been usually limited to 50 volts owing to weight and space considerations. A typical lay-out of long-wave naval receiving gear for a ship's office is indicated in Fig. 24, the plans in Figs. 5, 6 and 7 indicating the general lay-out of the receiving bays. Each bay is supplied from the common anode battery via a cushioning unit to prevent interference.

The question of adopting alternative power supply systems to the above accumulator systems has received considerable attention, and the earlier policy of adopting battery supplies entirely has undergone some revision. Experimental investigations are at present in hand with rectified alternating current and with d.c. generators as sources of supply (with or without a floating battery). If floating batteries are not used these systems must be capable of being operated from either the 220-volt d.c. or the auxiliary 20-volt ship's mains in order to provide for emergencies, but if floating batteries are used these in themselves form the emergency supply. The supply of anode voltage to receivers does not appear to present

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any insuperable difficulties, but the importance of good voltage regulation for a filament supply to dull-emitter valves in a central receiving room when the filament current may be changed suddenly from 5 amperes to a few milliamperes renders the problem much more difficult. It is essential to arrange for automatic voltage-control devices to be fitted in the 220-volt or 20-volt busbar supply to cover the wide variations in voltage which may occur in normal ship routine, whilst the inherent regulation of the motor-alternator or d.c. motor-generator must be very small. The best arrangement as regards regulation is probably a motor-generator system with floating battery, and in this case the floating battery is available for use alone in emergency; on the other hand the rectified a.c. system has considerable advantages, especially as regards reduction of interference.

The wiring and arrangements of the receivers to reduce interference has to be very carefully carried out; it may be interesting to enumerate here some of the chief interference problems which arise:—

(*a*) Mutual interaction from adjacent receivers, due to (i) common anode and filament supplies, (ii) local heterodynes.
(*b*) Morse-key clicks.
(*c*) Direct interference from transmitters adjacent to the receiving room.
(*d*) Interference between wireless telegraph offices in a ship, arising from mutual coupling between aerials.
(*e*) Faults such as earths or poor insulation on lighting supply leads, battery leads or fan leads introduce irregular low-frequency noises.
(*f*) Aerial down-hauls and stays which cross one another and rub are a serious cause of interference, especially to short-wave reception.
(*g*) Valve noises due to vibration, particularly in small ships.
(*h*) High-frequency interference from incoming cables which are connected to running machines.

The problem of interference from running machines is found to be especially troublesome, and in this connection the importance of good workmanship in the commutators and slip-rings of these machines in order to reduce sparking and ripple voltages cannot be overemphasized. Sparking at a commutator may radiate wireless waves directly to the central receiving room. In addition, sparking or machine ripple may set up high-frequency oscillations along the cables, and, if any of these have a direct conducting path to the central receiving room, serious interference to both long- and short-wave reception may result. For the case of machines with poor brushgear design the effect of brush chatter in setting up high-frequency interference in this manner has proved especially troublesome. It has been found necessary in certain cases to fit low-pass filter units not only in the output but also in the input, i.e motor circuits of machines, to reduce this trouble.

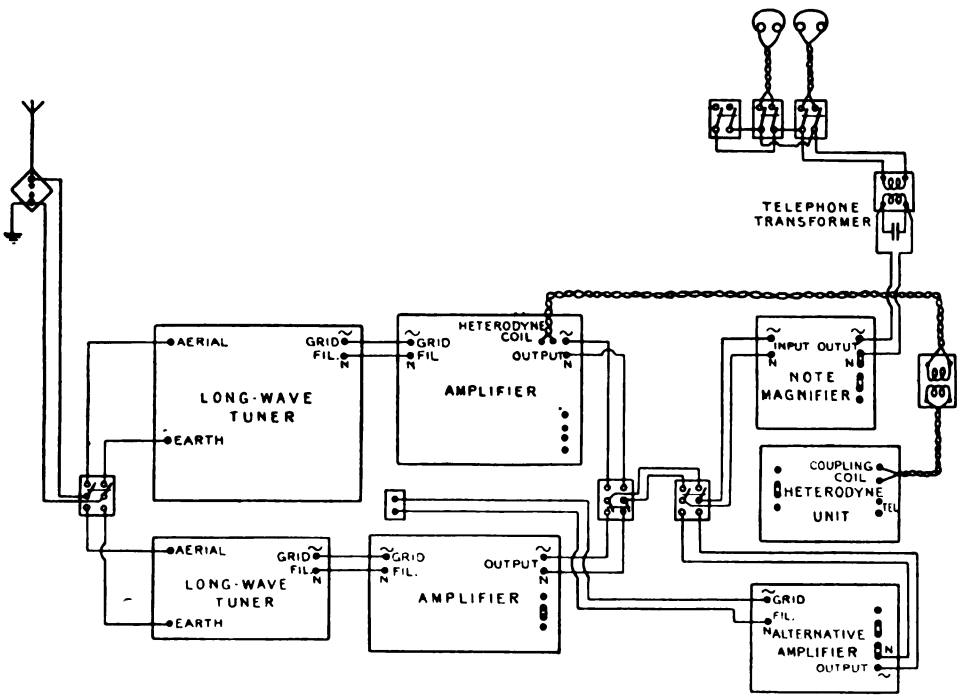

FIG. 24.—Arrangement of receiving model outfit.

(M.) *Receiving Aerials.*—The aerials employed for long-wave reception range from the main aerial of a ship to a short single wire. The aerial in use may be connected to the tuner in the office either by the central conductor of a long trunk or by a lead-cased cable which may have a length of the order of 150 ft. The effective capacity of the aerial receiving system may therefore

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vary over very wide limits, and the design of the aerial tuning circuits for a given receiver must be arranged to cover this wide variation. There is a considerable loss of receiving efficiency as the length of cable increases. More recently, cables having air and paper dielectric, to obtain selectivity and the inductive coupling can be loosened to any desired extent. In most cases, arrangements are made for listening-through on the wavelength being transmitted. For special purposes, such as direction-finding and submarine reception, where very

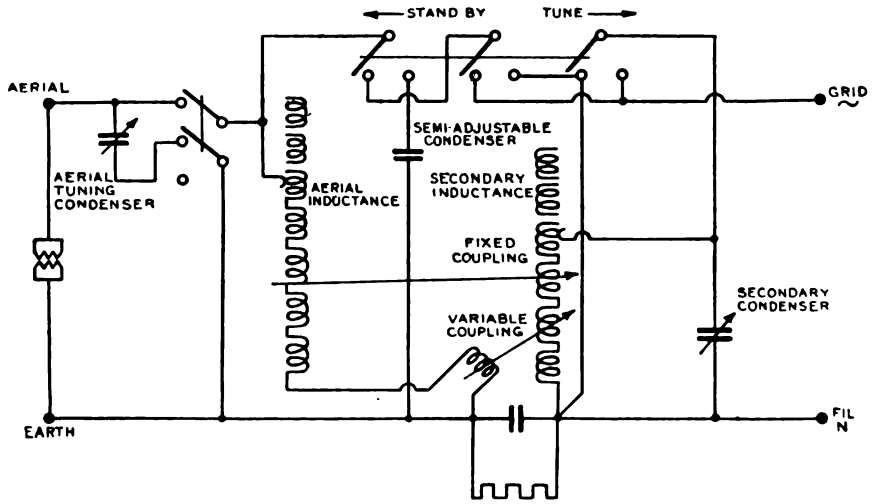

FIG. 25.—Schematic diagram of long-wave tuner.

and therefore much less capacity per unit length, have been employed with success in place of rubber cables.

(N.) *Receiver Design for Wide Frequency Bands.*

(i) *Long-wave receivers.*—Reception over the band of waves covering 12·5 to 6 000 kilocycles per sec. (24 000 to 50 m) has been provided by three ranges, viz. 12·5 to 600, 460 to 2 000 and 1 500 to 6 000 kilocycles per sec. (24 000 to 500, 650 to 150, and 200 to 50 m). The high amplification combined with very quick searching is desired, special high-magnification non-selective amplifiers have been developed. Also, special tuners have been developed to allow for the high damping and high capacity to earth of loop-aerial systems when used in submarines.

Figs. 25 and 26 illustrate the circuit arrangements of one of the naval long-wave tuners and amplifiers which has had considerable application. The tuner comprises

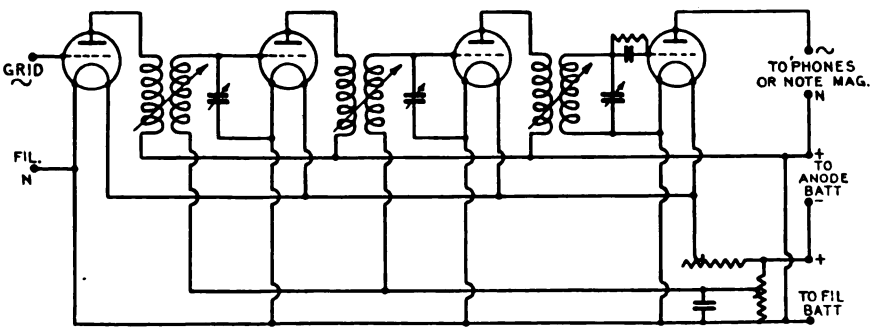

FIG. 26.—Schematic diagram of long-wave amplifier.

earlier policy was to adopt a unit form of assembly and construct a separate tuner and a high-frequency amplifier for each of these ranges. Also a standard 3-stage note-magnifier and a heterodyne to cover the range 12·5 to 2 500 kilocycles per sec. (24 000 to 120 m) was developed. Thus, any long-wave naval receiver included the appropriate tuner unit and high-frequency amplifier. The tuners are of the two-circuit magnetically-coupled type, and are fitted with a " stand-by " and " tune " switch. In the " stand-by " position the aerial circuit alone of the tuner is in use, so that quick searching is possible. In the " tune " position the closed secondary circuit is added two tuned magnetically-coupled circuits and is designed for

(i) tuning aerials having widely varying capacities for the entire range of 12·5 to 600 kilocycles per sec. (24 000 to 500 m), or
(ii) tuning as an acceptor circuit for the higher frequencies above 60 kilocycles per sec. (5 000 m).

A " stand-by " and " tune " switch is provided whereby the amplifier can be connected across the aerial tuning inductance when in the " stand-by " position, and across the secondary inductance when in the " tune "

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position. When in the " tune " position, a condenser having a capacity equal to that of the first valve of the amplifier and of the leads is connected across the primary inductance, thus obviating a change of tuning of the aerial circuit when switching over from " stand-by " to " tune." The secondary circuit is short-circuited and earthed in the " stand-by " position.

The amplifier circuits shown in Fig. 26 are designed for high amplification and selectivity. The entire range is covered by a number of range boxes which can be fitted into the amplifier. A separate heterodyne is employed which is coupled into the anode circuit of the first valve, thus to a great extent obviating radiation from the aerial.

For certain bands of waves, special provision has to be made for quick change from one wave to another, and this may be provided for either by the use of adjustable stops on the variable-condenser dial for

employing a detector valve operating on the threshold of oscillation in conjunction with a separate quenching valve. For automatic reception at shore stations, a special circuit has been developed using the double heterodyne principle. As regards short-wave reception in ships, the fact that reception in general has to be carried out via cables or aerial trunks introduces difficulties due to the fact that for certain waves these may tune as transmission lines. The effects of this may usually be greatly reduced by the use of very loose aerial coupling, although loss of signal strength is thereby occasioned, especially for the case of interrupted-continuous-wave signals.

A typical design of non-radiating short-wave receiver for ships is shown in Fig. 27. A screen-grid valve is employed in the first stage, followed by a detector valve with reaction which is used as an autodyne for continuous-wave reception. A separate note-magnifier is used

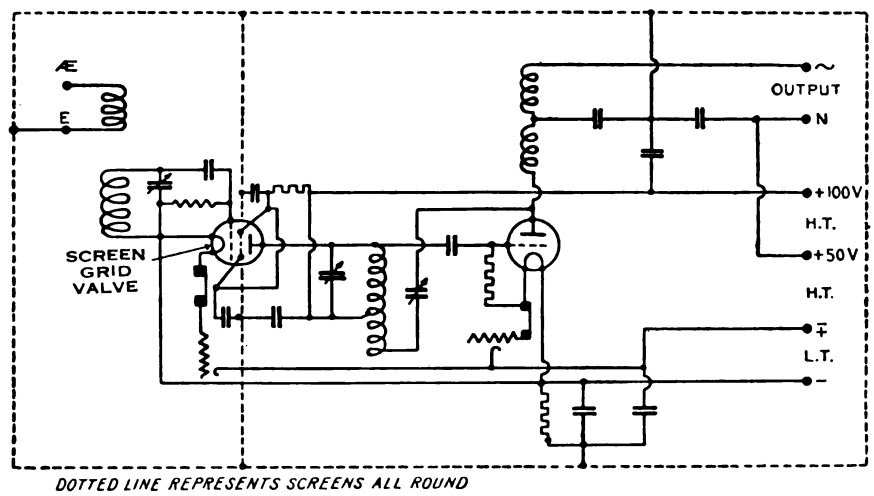

FIG. 27.—Short-wave tuner amplifier.

cases where the change is small, or by change-over from one circuit to another for cases where there is a large difference between the two waves. The requirement that the majority of naval receivers must be non-radiating, necessitates an increase of circuit complications and in many cases increases the number of valves required for a given amplification.

More recently the general policy of the design of receiving gear has been modified, the unit construction being departed from to a considerable extent. The later designs of amplifiers now include note-magnification and separate heterodyne in one unit, and in the case of reception on longer waves of the order of 600 m and above, special provision is made for multiple reception on one aerial.

(ii) *Short-wave receivers.*—To cover the short-wave requirements, two types of special tuner-amplifiers have been developed, one a simple oscillating detector with two stages of note-magnification and no high-frequency amplification, the other an oscillatory detector preceded by a stage of high-frequency amplification to prevent oscillations from the detector from reaching the aerial and being radiated. Also, receivers have been developed behind this model. Despite the use of the 4-electrode valve, there was found to be coupling between the tuned grid and tuned anode circuits. This coupling was reduced by introducing a very small condenser between the tuned grid circuit and the grid of the valve.

(O.) *Receiver Components and Assembly.*—The reliability in performance of receiving instruments for naval use is naturally of first importance. The apparatus has to be used in all climates, and, whilst every effort is made to reduce size and weight, instruments designed for use in H.M. ships are generally of a more substantial construction than is considered necessary for shore work.

A number of parts, components and units which lend themselves to assembly in various ways for various purposes have been developed. These are embodied in the instrument designs, with the result that certain standard practices tend to develop for naval work, with a consequent saving in cost and in the time required for production. Where the quantities required are large, mouldings, die castings and stampings are used in these parts, the necessary stock being held by the Admiralty and issued to contractors as required. In addition, commercially produced components which are stan-

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dardized in the trade and suitable for the required purpose are used whenever possible.

The adaptable receiving components specially designed include an adjustable condenser with die-cast fixed and moving vane units and with moulded end plates, a series of dial switches and a series of barrel-type switches each capable of assembly for many varied purposes, filament rheostats and potentiometers, a slow-motion dial, grid leak and anode-resistance holders.

The adjustable condenser can be assembled to either 0·5, 1·0, 1·5 or 2·0 jars' * maximum capacity. It is of the twin pattern in which two sets of moving vanes, insulated from one another and mounted on opposite sides of the spindle and in the same planes, move between similarly mounted fixed vanes. Each set of moving vanes is connected to one of the sets of fixed vanes. This construction is well known and gives the maximum capacity change in a given space. The vanes are stamped from sea-resisting aluminium alloy and are held together in forms die-cast on to them. The units are assembled between end plates moulded from insulating material. These end plates carry the bushes for the spindle. The moving system is capable of adjustment endways for centring. The whole provides a strong and rigid construction capable of being assembled to a high value of maximum capacity for the space occupied, with a mechanically balanced moving system. The minimum capacity is of a fairly low order (0·02 to 0·05 jar). These condensers are not particularly suited for short-wave receiving work, and other types are being developed for this purpose.

The dial-type switches are intended for mounting directly on the faces of instruments, and the contacts and brushes are exposed. The barrel-type switches are mounted behind the instrument front. They are self-insulated and can be mounted on metal panels. They take up much less panel space than the dial type. Both types of switch can be built for many purposes.

The slow-motion dial is worm-geared with a micrometer thimble on the worm shaft. The worm is held in close mesh with the wheel by a spring, and provision is made to enable it to be immediately disengaged to permit quick setting. The design is primarily intended for wavemeters and receiving instruments requiring a high accuracy of adjustment.

A vertical and a horizontal anti-vibration 4-pin valve-holder have also been designed specially to suit service requirements. The features of these are that provision is made for locking-in the valve, damp and dust are excluded and the connections are enclosed. The valve-socket block is carried on rubber to provide an effective cushion with damping. The holders are designed to enable the rubber cushion to be replaced from the front and without disconnecting the valve-holder.

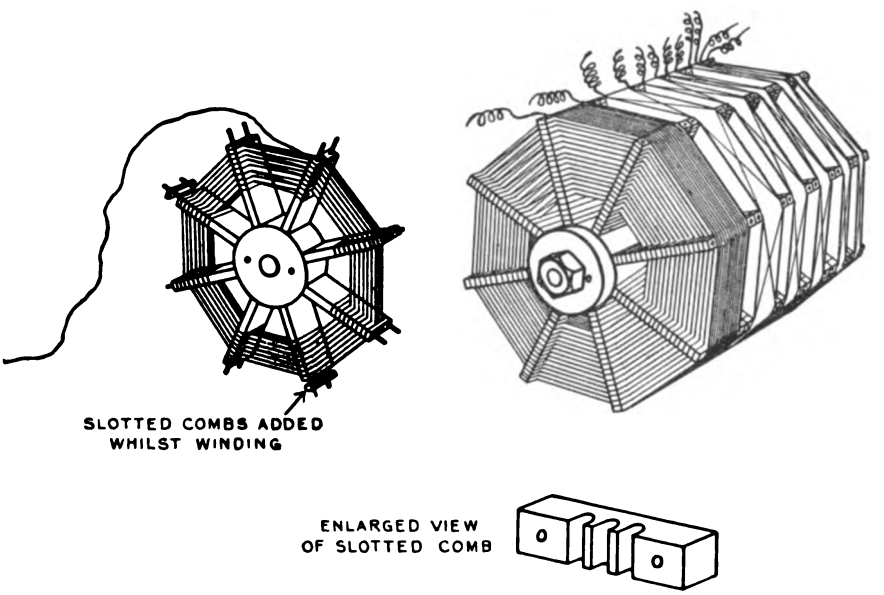

FIG. 28.

As regards the inductance coils for long-wave tuners, an endeavour is made to obtain efficient coils with a minimum of self-capacity. The theoretical maximum efficiency cannot, in general, be obtained, however, owing to limitations imposed by the permissible size of the instrument and by manufacturing conditions. In general, all coils are space-wound, one form of winding which has been adopted being shown in Fig. 28. The wire used in this case is 9/36 S.W.G., each strand being enamelled and silk-covered. The use of enamelled wire, each strand silk-covered, was found in practice to give a higher efficiency than the use of enamelled wire only due to the elimination of broken strands with the additional silk covering. Smaller coils are, in general, space-wound with 9/36 S.W.G. wire on threaded formers for ease of manufacture. When very short wavelengths are in question, losses in the former become of more importance than those in the wire. Hence solid wire is then used. For naval purposes the use of coils entirely

* 1 jar = 1 000 cm.

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without solid formers is impracticable, and a compromise has been reached by the use of ribbed formers, the wire being held in threads cut in the ribs. Thus, each turn of wire is supported in air and only touches the actual ebonite former by ribs on the latter. This construction has been found to give satisfaction since, in addition to reducing losses generally, it forms a means of obviating the difficulties due to the formers "sweating" in hot climates.

Receiving instruments are mounted on bulkheads, and the space available is very limited. Consequently, controls must be on the face of the instrument, and it becomes an advantage that all parts shall come away with the instrument front. For this reason it is usual to provide a removable reversible cover which protects the front while the instrument is in transit and also serves as a bracket for securing it to the bulkhead when in use. In the later designs, the box and cover are of cast sea-resisting aluminium alloy, and the instrument front is of a similar rolled alloy.

It may be of interest to describe a small point in the design of variable inductances and condensers. Naval apparatus has hitherto been so designed that right-hand rotation gives an increased scale reading and indicates an increase of capacity or inductance and wavelength. In accordance with the decisions reached at the Washington Radio Convention the Navy has now adopted the use of kilocycles and consequently for the above apparatus a right-hand rotation corresponds to a decrease of frequency. For present and future developments of apparatus it has now been decided that right-hand rotation, corresponding to increase of scale reading, shall give increase of frequency. Consequently it has been necessary in a number of cases so to re-design the movements of condensers and inductances that right-hand or clockwise rotation gives a decrease in capacity or inductance.

(P.) *Direction-finding.*—The application of wireless telegraphy to direction-finding has great tactical importance. We may summarize the principal uses of direction-finding as:—

(1) Ascertaining the direction of enemy ships or aircraft.
(2) Checking the position of our own ships or aeroplanes making reports.
(3) Assisting aeroplanes in reconnaissances.
(4) An aid to navigation.

The type of system which is fitted in a particular ship largely depends on the structure of the ship. Just as a magnetic compass placed on board a steel ship is liable to considerable errors due to the magnetism of the ship (which errors have to be suitably corrected), so also a direction-finder is affected by the ship's structures and needs adjustment and calibration before reliable bearings can be obtained, and special correction devices have to be fitted for this purpose as with a compass.*

As a general rule it is the practice to fit the Bellini-Tosi system consisting of two loops in planes at right angles. Sometimes, however, where no suitable loops can be rigged, a rotating frame coil is fitted. Experience has shown that for a frame coil to be a success it must be placed in a position very well clear of metal work and preferably as high up as possible. In order that the coil may be conveniently rotated, the operator and instruments must be housed quite near to the coil, and sometimes this may limit very seriously the positions possible for the coil.

The Bellini-Tosi system is always fitted with one loop in the fore-and-aft line and the other loop exactly athwartships. To get the best results, regard must be paid to the positions of all large masses of metal in the vicinity of the loops, and generally the aim is so to place the loops as to secure the greatest possible symmetry with respect to the whole metal work of the ship. The loop run fore and aft is the one which invariably gives the greatest trouble in this respect.

The best position for the office, from a purely wireless point of view, is immediately beneath the aerials, but attention must be paid to other factors, such as weight distribution, vulnerability and compactness. For these reasons it is often necessary to have the instruments at a considerable distance from the aerials to which they are connected by cables. A common position for the operator and instruments is either the main or auxiliary wireless office. For the shorter waves, however, the price paid as regards accuracy for this convenience of organization and supervision is heavy, and offices immediately beneath the aerials are capable of much better accuracy.

(Q.) *Direction-finding Instrumental Design.*— The primary factors which govern the design of the receiving instruments for use in direction-finding are:—

(i) True bearings must be obtained without elaborate corrections or calculations.
(ii) Considerable amplification.
(iii) Ability to search rapidly over a band of waves so that a rough bearing may be obtained even though a transmission may last only a few seconds.

To meet these difficult requirements as far as possible, the moving element of the goniometer or frame coil is made as light as practicable to increase the speed at which a bearing can be taken, and generally the receivers have an unselective position suitable for quick search and a much more selective position for accurate work on a given band of waves.

The directional determinations are usually made by ascertaining the angle at which the search coil in a goniometer or a frame coil has to be set to give the minimum strength for the incoming signal. It is usual to find this angle by taking two readings at equal strengths, one on each side of the position of minimum strength, and to deduce the mean of these readings. To facilitate the determination of the mean value and to eliminate calculation, the naval service radiogoniometers and training units for frame coils are fitted with an automatic angle-dividing device. An additional rotatable scale is also provided which is driven by, and held in step with, the master gyro-compass. Thus, a bearing relative to the ship's head can be taken directly on the fixed

* C. E. Horton: "Wireless Direction-Finding in Steel Ships," *Journal I.E.E.*, 1923, vol. 61, p. 1049.

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scale, and true bearings are indicated directly on the rotating scale, no calculation being required for either. These two refinements, introduced into naval designs several years ago, greatly reduce both the time required to take a bearing and the error.

The ordinary direction-finder does not differentiate between the true bearing and one differing from it by 180°. For tactical purposes it is essential to be able to determine the absolute direction, and "sense finders" are fitted for this purpose. In the case of the Bellini-Tosi system of loops the sense finder is a separate unit attached to the receiver and is not in use when an accurate bearing is being taken. It is subsequently switched into action to determine which of the two readings 180° apart is the correct one. Separate sense-finding aerials are not normally used but sometimes when the office is at a great distance from the Bellini-Tosi loops an auxiliary aerial of the open type is necessary; this is usually a single vertical wire. For the case of a rotating coil or a system of fixed coils it is usual to have a separate sense-finding aerial, and the sense finder is incorporated as a part of the receiver.

Part (4). Organization and Development Procedure.

The responsibility for the experimental development and adaptation of the apparatus for naval communications rests with the Captain, H.M. Signal School, Portsmouth, where a staff of naval signal and civilian officers work side by side. The naval officers are attached to the Signal School for short periods only, usually of the order of two years, and in this way there is a continuous flow of officers from ships at sea who have intimate and up-to-date knowledge of the performance of apparatus in use or under trial at sea. These officers are able to state the broad requirements to which new apparatus should conform for success under sea-going conditions, and they co-operate with the civilian officers to ensure that these requirements are fulfilled.

The Director of the Signal Department, Admiralty, is in the closest touch in all signal matters with the Captain, H.M. Signal School, and is in direct touch with the Board of Admiralty; the reports from the various Commanders-in-Chief at sea covering wireless telegraph apparatus and naval communications are considered by the Director of the Signal Department, and the policies to be adopted as a result of these reports are formulated by him, after consultation with the Captain of H.M. Signal School, for the approval of the Board of Admiralty. In this way it is ensured that the experimental work of the Signal School is in accordance with the most recent sea-going requirements.

The experimental scientific staff of the Signal School are pooled under the Director of Scientific Research for the Admiralty, whose sphere includes all matters which relate to the supply and general distribution of the scientific civilian staffs for naval experimental and research purposes.

A close liaison exists between the Signal School and the other Government and industrial organizations, such as the Radio Research Board, the National Physical Laboratory and the British Engineering Standards Association. Representatives for the Admiralty are appointed from H.M. Signal School to various committees dealing with the work covered by the above organizations. The contact with the latest experimental and research work, dealing especially with wireless, has proved of great value to the Signal School.

Apart from the relations which naturally exist between the various scientific staffs of the Admiralty which are lent by the Director of Scientific Research to various naval establishments, a close liaison with the other military services is maintained by means of the main Wireless Telegraph Board Committee and the experimental sub-committee of this Board. These committees meet at frequent intervals and discuss service experimental developments. In addition, regular visits are interchanged with the service experimental establishments and arrangements are also made whereby, as far as practicable, components are standardized for inter-service use.

As stated above, the responsibility for development and design of apparatus in accordance with Admiralty policy is vested in the Captain, H.M. Signal School. Following on the production and satisfactory tests of first experimental models, first in the Signal School laboratories and then at sea, the final designs and specifications for the components are prepared and competitive tenders invited. The apparatus delivered from firms is tested for acceptance by the Signal School, and is later installed in ships under the direction of Signal School fitting-out officers.

In preparing the final designs of apparatus, commercial methods of production and workshop practice are kept in view, and every effort is made to confine the demand to common processes. At the same time, a high standard of workmanship is invariably called for in the specifications to ensure reliability of operation under sea-going conditions. B.E.S.A. standards are always adopted whenever possible, but proprietary articles are not usually embodied in designs for Service use unless they are of sufficient outstanding merit to out-balance the disadvantages of dependence on any particular source of supply.

Finally the authors desire to emphasize the debt owed, not only to members of the Service establishments but to manufacturers of equipment, who by their cordial co-operation have contributed to the ideal which all have in mind, that is the efficiency of the British Navy.

The paper is published with the permission of the Board of Admiralty. The authors desire to thank the Captain, H.M. Signal School, Portsmouth, for assistance in the preparation of the paper and for the use of diagrams prepared for Admiralty wireless telegraph handbooks. In addition, they would express their indebtedness to the members of the Civilian and Naval Experimental Staff of H.M. Signal School who have assisted in the preparation of the paper.

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Discussion before the Wireless Section, 4th December, 1929.

Commander J. A. Slee: As the paper suggests, wireless communication in the Navy started with what is called a " plain aerial outfit," with which it was very important to maintain the standard of insulation in the aerial, which consisted of a length of wire very heavily insulated with rubber; I think the rubber was actually put on by hand. Next we used a circuit coupled between the transmitting gear and the aerials. There were two separate transmitters and two separate receivers making two separate sets of different wavelengths, which were then called A and B " tunes." There was no official transmitting apparatus in those days; we had a complete and properly constructed receiver—of the coherer type, of course—but we had to build up our own transmitters to specifications supplied, and to arrange with the carpenters to build various wooden formers, in the construction of which no metal was to be employed. Rubber-covered wire was used. There was very keen competition between ship and ship. Attempts were made to use alternators instead of batteries, employing the coil as a transformer. We had no alternators and very few ventilating fans, the latter being regarded as a great innovation at that time. Then followed what was called C " tune," which was a set of apparatus, not a specific wavelength, and we considered this to be a great advance. We used a 2½-kW motor, and, with an open spark-gap, the spark passed with a very loud report. With the object of increasing ranges, we tried flying a kite at the end of a long piece of wire over the end of the aerial, and obtained some extraordinary results which, however, did not last for very long, because the kites were not really strong enough for the purpose. The early coherer receivers were followed by magnetic detectors, and that marked the beginning of audible reception. In those far-off days there was no telegraphist branch in the Navy: the torpedo officer was placed in charge of the wireless apparatus. After the magnetic detectors, crystal receivers were introduced. The first crystal used was a dull red stone, named zincite. Then came electrolytic detectors, which we constructed for ourselves; we made up half a dozen at a time and tried them against one another. I think we can all feel a considerable amount of pride in the state of efficiency to which wireless communication in the Navy has been raised since the days of which I have been speaking.

Prof. C. L. Fortescue: For those members who have been in any way connected with the Naval Service the paper has an interest from the historical point of view, but to the many who have not been intimately in touch with the development of naval wireless telegraphy the interest is in the insight which the paper gives into the conditions under which naval wireless is carried on. These conditions are very different from any found elsewhere. For instance, the paper describes cases of single ships in which six or eight lines of communication are all operating simultaneously. Then, again, many special difficulties are encountered—for example, the effect of sea-spray, or a heavy coating of salt over the insulators —which have not generally to be provided for in ordinary wireless installations. The skill called for in the design of the trunk leads on board ship is demonstrated in Fig. 23, and very few designers of apparatus for wireless reception are prepared to make provision for mounting their instruments on a deck which, as a result of gunfire, may whip a matter of about ½ in.! It is not surprising that the apparatus has the external appearance of being complicated, because so much has to be crowded into a small space. The master oscillator, with its exact electrical balance all the way through, is interesting.

Commander C. S. Daniel: Naval wireless-telegraph work has certain definite characteristics. Point-to-point working is by no means the most important; the most important communications are those of the main fleet, which require absolute certainty of interception and wireless discipline. In making enemy reports, the reporting ship or aeroplane obtains an answer from only one other ship or aeroplane. The rest of the fleet must rely on interception. Further, on certain lines, the fleet flagship must be able to stop one ship, direct some others to carry on, and generally control traffic. This requires most efficient listening-through and good discipline. Given this, it should be possible to pass a short manœuvring signal to a fleet of 20 to 30 ships, obtain the answers, unit by unit, and be ready to make the executive signal in not more than 1½ to 2 minutes. Obviously, naval apparatus must be so designed that sets will keep as constant a frequency as possible; and the most accurate wavemeters must be provided in order to avoid tuning discrepancies, which would be fatal to interception. A difficulty arises here, as ships may have to change their wavelength in order to avoid jamming by the enemy, or if there is congestion in their own fleet. The wavelength to which they change must be accurate. In shore-station practice we could provide as many sets as there were wavelengths required, but, as pointed out in the paper, space on board ship is very limited. The design of naval wireless-telegraph apparatus is far from easy. What degree of constancy can be obtained in a set which has to change wavelength almost instantaneously, and what developments are taking place to improve this? It may sometimes appear that naval specifications are very exacting; this, however, is bound to be the case. The importance of the battle of communications which precedes the fleet action cannot be over-estimated. Any defect in the wireless apparatus on board a ship of the fleet may alter the course of a war. The various things which *may* go wrong are very numerous: e.g. an earthed aerial, a burnt-out filament, some tuning error in transmitter or receiver, etc. Obviously, many features in naval design are dictated by experience afloat. Reference has been made in the paper to the new graduations for condensers and inductances, i.e. decrease of capacity or inductance with increase of scale reading. At present an increase of scale reading implies an increase in capacity or inductance. In the fleet all reference to waves is now in terms of kilocycles per sec.; and nothing is more irritating, when a fleet is being tuned in, than the fact that operators must decrease their scale readings when ordered to

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increase their frequency. Referring to rotating-coil direction-finders, the paper states that the operator should be near the coil and that this imposes certain limitations. It is hoped to overcome these limitations by leaving the coil up above, putting the operator and his gear below, and rotating the coil by remote control. At the beginning of the paper the authors mentioned the good results which accrued from collaboration between the scientist and the naval officer. I should like to support that statement whole-heartedly. The naval officer has to take the responsibility for communications at sea; he has certain apparatus to do it with and he must get the best results from that apparatus. I am glad to have this opportunity of expressing my admiration of the way in which the scientist meets the very exacting demands of the naval officer. So long as this co-operation exists, I think that naval wireless-telegraph communications will not be found wanting.

Dr. C. V. Drysdale: As Commander Slee and Prof. Fortescue have remarked, the difficulties associated with naval wireless work are great, and this ought to be fully recognized when comparing some of the devices with those in general use. The problem of supporting the valves against the severe shocks of gunfire and yet to permit of their rapid replacement is by no means an easy one; the need for working with a number of wavelengths and rapid change-over from one to another introduces difficulties which will be readily appreciated; and the proximity of large metal structures and the necessity for bringing the leads through trunks and glands add seriously to those difficulties. Accurate direction-finding is obviously of great importance to the Navy, and here the effects of the hull, superstructure and stays are most in evidence; but, by careful choice of positions and compensation, accuracy has been secured. On the scientific side the existence of a large fleet under a single organization and working over such a large range of frequencies has enabled much valuable information to be obtained on transmission under varying conditions. Such information is bound in time to have useful effect not only on naval but upon general navigational and commercial wireless, and we may be grateful to the officers and the scientific and technical staff of the Navy for the great advances they have made.

Mr. L. B. Turner: From this interesting description of past and present naval practice and apparatus I have noticed what is, I think, a rather remarkable omission: I see no reference to telephony. At first I thought this might be because the telephone, for strategic or other reasons, is debarred in the Naval Service; but the reference in Capt. Dorling's opening remarks to the Nelson of the future putting a telephone to his blind ear suggests that this is not so. The value of American whitewood in high-frequency fields was the discovery, I think, of the Post Office engineers in their work in designing the Rugby inductances. It is pointed out that its good qualities are destroyed by a humid atmosphere. Has impregnation of any kind been tried? Paraffin-wax suggests itself, but probably this is ruled out owing to the danger of fire. My experience with cellulose acetate in acetone suggests that this might be a suitable material to apply. The authors urge us not to forget that there is an optimum spacing between the strands of the high-frequency cable, and that the greatest possible number of strands in a given space is not the best. I should like to know how the correct spacing is provided, i.e. whether there is a whipping of spacing material on the conductor, and what material is used. Presumably cotton is unsuitable because of the enormous losses introduced unless it is very dry. Can the authors give us some details of a typical piece of one of these cables? The $\sqrt{N}$ rule for tapping a coil of N turns, given on page 22, I find novel and pleasing. An interesting point arises in connection with Fig. 21, showing the frequency-doubling circuits. I notice that there is no d.c. supply to the anodes. With back-coupling present, operation without d.c. supply seems possible, but does Fig. 21 represent actual conditions? On page 22 it is stated that the power factor of a good mica condenser is of the order of 0·4 of 1 per cent. This seems to be a very high figure. Good-quality mica condensers in laboratory work have very much lower power factors than this; and the Dubilier condensers at Rugby are believed to show a power factor of 0·02 of 1 per cent.* Fig. 26 shows an amplifier having 3 high-frequency stages, with a wavelength range of 24 000 to 500 m, and yet without any sort of neutrodyne retroaction balance. This seems to be a most remarkable receiver, and it would be very interesting if the authors could tell us a little more about it.

Mr. F. S. Barton: In regard to the re-design of condensers and inductances so that a clockwise movement gives a decrease of inductance or capacity, I should prefer to see standard practice retained and the scales arranged to read anti-clockwise for increasing frequency. This would have the advantage of still retaining the use of conventional commercial components, in which, as at present, a right-hand or clockwise motion gives an increase in the quantity being varied.

Mr. F. C. Lunnon: One problem in the case of a ship short-wave transmitter to cover a large range of wavelengths is that of providing suitable coupling arrangements between the output circuit of the transmitter and the aerial, so that at every wavelength one can transfer sufficient energy to the aerial fully to load the transmitter. The difficulty arises because in a ship one is forced to use a fixed length of aerial for the complete range of wavelengths. With a reasonable length of wire clear of the deck insulator, plus, in many cases, a considerable length of wire through an aerial trunk from the deck to the wireless cabin, the total length of aerial, measured from the transmitter, will usually be several wavelengths at the shortest wavelength and more than $\frac{1}{4}$ wavelength at the longest wavelength. It follows that the correct termination for the aerial varies rapidly over wide limits between the shortest and longest wavelengths of the transmitter, and unless means are provided for approximating to the correct termination then, however much mutual inductance be provided between transmitter and aerial, there will be many wavelengths at which only a very small amount of energy can be transferred to the aerial. When the total length of aerial is a whole number of $\frac{1}{2}$ wavelengths the correct termination is given by a tuned parallel circuit of inductance and capacity; and for an odd number of

* See *Journal I.E.E.*, 1927, vol. 65, p. 297.

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¼ wavelengths a tuned series circuit is required. Moreover, the required values of capacity and inductance will themselves vary greatly over a range of, say, 14 to 70 m. I have found it necessary, for such a wave-band, to provide two variable condensers—one very small—and two equal inductances. Switches are provided to select either variable condenser and connect it in parallel, or in series, with the two inductances, which themselves may be switched in series or parallel. The latter provide the mutual inductance between transmitter and aerial. Even with these complicated arrangements the aerial, more often than not, is a reactive impedance requiring a large mutual inductance to transfer sufficient energy fully to load the transmitter. I was therefore surprised to note, from Fig. 19, that the coupling to the aerial for the short-wave transmitter, 14 to 70 m, is shown as a simple coupling coil. I should like to ask the authors whether in actual fact that is what is employed and whether, in that case, they do not experience many points in the wave-range where only a very small amount of energy can be transferred to the aerial. In ships where a very long aerial trunk is necessary between a short-wave transmitter and the deck, I should like to ask the authors whether the conductor through the trunk is used as a continuation of the aerial or as a feeder to the point where the aerial emerges from the trunk. The former makes for simple wave-changing but involves considerable losses at short wavelengths; the latter, of course, is a more efficient method but involves remote control of terminating apparatus located at the deck end of the trunk. I notice that reference is made to the extensive use of ebonite as an insulating material, both on receivers and transmitters. General experience of ebonite in tropical climates is that both its insulating and mechanical qualities deteriorate greatly. Mycalex, on the other hand, is free from both these defects. What is the authors' experience of ebonite in ships stationed for long periods in the tropics, and have they found mycalex generally satisfactory as a substitute for ebonite?

Mr. H. Faulkner: It seems to me at first sight that the protective arrangements provided on the transmitter are inadequate. There appear to be no overload relays provided on the switches, such as one would expect to see. The protective arrangements for the oscillating valves themselves consist simply of a single relay to protect the 3 valves which are connected in parallel. While I agree that such protection is adequate as protection, it is my experience that this arrangement is very often likely to cause needless delay because of the difficulty in tracing which of a number of valves has failed. My experience, perhaps, has been more concerned with cooled-anode valves, and the circumstances may be different; but certainly I would suggest that the fitting of a small indicator, if a relay cannot be fitted, to indicate which valve has taken an overload, would be an improvement. I have no doubt that valves do *sometimes* fail from loss of vacuum. I notice that horn-gap fuses are fitted in the rectifier. Are these the principal protective devices for the system in the case of a breakdown of the smoothing condenser, for instance, or other short-circuit on the supply, or is any other form of protection provided? If these are the only protective devices, I am interested to know whether they are completely satisfactory, and what kind of fuse wire is used. With regard to keying, I note that in Fig. 11 this is direct in the anode circuit. This seems to me rather a "hard" method of keying. A great deal of sparking must take place at the key when in that position in the circuit, and it would be practically impossible to do any high-speed work. Some time ago I tried a system of keying in the grid circuit and a.c. supply simultaneously, with the idea that the breaking of the grid circuit stopped the current, and the breaking of the a. c. supply prevented the undue rise of anode voltage from the rectifier through taking off the load. Some improvement of the sparking at the key might be obtained by such an arrangement. Further, no coupled circuit to the antennæ is shown; is this due to a lack of space, or to the fact that it is not found necessary? I should have thought that with large numbers of transmitters working on the same ship, and on a large number of ships comparatively close together working on widely different wavelengths, there would be considerable trouble from harmonic radiation. Referring to Fig. 13, it is observed that each valve is provided with its own grid leak. Have the authors any evidence as to whether this arrangement equalizes the load between the valves? Post Office practice has always been to provide a common grid-leak for the circuit. I do not think that individual grid-leaks can have any effect on possible spurious oscillations between the valves. In Fig. 14 the overload contacts on the anode relay, etc., are shown as opening the filament circuit—that is to say, when an overload occurs the filament of the valve is switched off. I should like to know whether that is found to be satisfactory. It is my experience that when a valve is very soft the overload may continue through the valve, even if the filament *is* switched off. With regard to Figs. 19 and 20, it seems remarkable that the arrangement provided for switching from short to long waves does not make the transmitter very unstable on the short waves. I should like to hear the authors' experience in this connection. I think that the frequency-doubler circuit should be made rather clearer in the diagram. What is the efficiency of this doubler circuit? I notice that the doubling takes place at high power; hence it would appear that it must be very effective, since otherwise much waste of power would take place.

Mr. G. Shearing and **Captain J. W. S. Dorling** (*in reply*): Commander Slee has given additional information relating to the earlier history of naval wireless communications and to some of the difficulties which were encountered in those days. One of the objects of the paper was to give a broad view of the requirements for naval wireless telegraphy apparatus and we are glad to note the remarks of Prof. Fortescue and of Dr. Drysdale in this connection. It is true that the varied requirements tend to make the apparatus look complicated; the limited space which is available for these requirements, particularly the requirement of the flexibility of sets over a wide wavelength range, necessitates the use of somewhat simple circuits. It is for this reason that coupled circuits for long-wave working are not usually employed in the high-power transmitters.

The necessity for rapid valve-change referred to by Dr. Drysdale has had to receive special attention, and

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the later designs of valve-holders for high-power sets enable any single valve to be replaced in a few seconds. Incidentally, the naval silica valve, which is largely employed for power transmitters, lends itself readily to replacement in the minimum of time.

The special requirements for naval ship working have been amplified by Commander Daniel, and these requirements emphasize the necessity for simple oscillatory-circuit apparatus. On the other hand, the control and operating circuits have to be fairly complex to ensure that the operations necessary for wave-change and for bringing various sets into use may be effected in the small time available for operating and manœuvring requirements. With regard to the need for extreme accuracy in the tuning of transmitters and the degree of constancy obtainable in a set which has to change wave almost instantaneously, the question can only be answered in general terms. The spark-transmitters formerly in use were tuned with carefully-calibrated wavemeters by skilled personnel, and no difficulty was experienced, since tuning was necessarily broad. The introduction of the valve transmitters for continuous waves or interrupted continuous waves, and the use of short waves, have naturally increased the difficulty of ensuring that all ships can transmit sufficiently accurately on any one of a number of waves. Wavemeters accurate enough for spark-transmitters lacked the necessary high degree of accuracy, particularly after extended use at sea. As a temporary expedient it has become necessary for one ship of the fleet, usually the flagship, to tune the transmissions of all ships to the same frequency; each ship, having noted the adjustments, can radiate the same frequency on subsequent occasions with reasonable accuracy. Improvements in wavemeters and the introduction of master-drive transmitters are leading to greater accuracy of initial tuning, although it is not yet certain that any substandard wavemeter of reasonable size, weight and simplicity, will be able to detect a small change of frequency as readily as can a receiver at a distance.

Mr. Turner has remarked upon the lack of reference to telephony in the paper. At one time it was thought that there would be great possibilities for wireless telephony in the naval service. Trial afloat showed, however, that the advantages were more apparent than real, except for certain definite applications. The main disadvantage is the lack of secrecy. Whatever system is used there can, at present, never be certainty that conversations cannot be overheard. Moreover, there can be no written record for future reference. The time taken to speak a message is often greater than would be sufficient to transmit the same message by morse, employing code, and the range of communication by telephony is less than the telegraph range. Again, even for the case when the morse is plain language the time taken is often less than for wireless telephony. Using wireless telephony, speech must be slow in order to avoid phonetic errors and repetitions. The use of wireless telephony in the naval service is therefore confined to those cases where secrecy is not essential and where direct communication between principals is required without intermediate links, or where conditions are such that it is not possible to arrange for operators trained in the Morse Code. With regard to American whitewood, impregnation with various bakelite compounds has been tried, but these have been found to be unsatisfactory. Experimental tests have been made of woods treated with paraffin-wax, also with cellulose-acetate in acetone, but the danger of fire has precluded their use. The heating adjacent to high-potential metallic contacts causes the solvent to give off a gas which is mainly hydrogen. This gas has a lower ionization-potential than air and therefore the danger of breakdown is increased. The experiments with paraffin-wax were definitely unsatisfactory. Also the addition of any form of impregnation to the wood provides an increase of permittivity and enhances the losses adjacent to high-potential metal contacts carried by the wood. An occasional flash-over has been found to destroy the insulation of wood so that replacement is necessary. In the case of ebonite, however, an occasional flash-over can usually be rectified by scraping away the burnt surface. A full discussion of the spacing of turns of stranded wire on a coil and of the effects of the size and spacing of the individual strands in the stranded wire has been published.* Although the effects of capacity between the strands and turns of a coil are not taken into account, the calculations give a valuable indication of the coil resistance to be expected. The size of the strand is usually fixed by manufacturing limitations. The spacing between the separate conductors of a strand is effected by enamel alone or by enamel together with a silk covering. The size of enamelled wire generally used is No. 38 S.W.G. When enamelling is employed alone, the separate conductors are liable to stretch, harden, and break during stranding and, in addition, the enamel may be rubbed at places so that contact with other strands is possible. Despite these disadvantages, large quantities of wire thus insulated are still used. For cases where it is necessary to ensure that the separate conductors are not broken, such as in the case of receiving coils using a smaller number of strands, No. 36 S.W.G. wire is used and each conductor is enamelled and then silk-covered. This affects the overall diameter of the wire and the spacing of the strands; difficulties are often encountered due to limitations of the space available for coils and for the tapping-points required. These limitations in many cases largely determine the design so that the actual resistance of the coil may have to take a secondary place, although its value is calculated as a check. Typical designs of stranded wire are indicated below:—

(*a*) 729/38 *Stranded Wire for Transmitter Coils.*—This is built up by stranding No. 38 S.W.G. enamel-insulated copper wires in groups of three (and continuing the operation). The lay of the groups should be right- and left-hand alternately. The stranded wire is covered with cotton braid which is coated with Ohmaline or Ozokerite. When assembling the wire without braided covering it was found that some of the wires snapped when bent over the combs in the coil-frameworks. These broken wires may cause brushing and insulation failure. The braiding overcomes this difficulty and provides a soft bed when bending the wire; also

* S. Butterworth: "The Effective Resistance of Inductance Coils at Radio Frequency," *Experimental Wireless and Wireless Engineer*, 1926, vol. 3, pp. 203, 309, 417 and 483.

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any broken strands are retained in place. In a few special cases transmitting-coils have been wound using wires with each strand silk-covered, but the cost of this is normally not considered to be justified.

(*b*) 9/36 *Stranded Wire.*—This wire is used in coils for receiving apparatus. It is stranded as described above, but each strand is enamelled and silk-covered. The chief function of the silk covering is to provide a soft bed for the wires when building-up during stranding and thus to prevent elongation, hardening and breaking.

We agree that the value quoted for the power factor of a good mica condenser is too high; this value should have been 0·2 of 1 per cent, to which value the figure has been corrected.

With regard to the amplifier and tuner shown in Figs. 25 and 26, these instruments were designed several years ago and have had a very wide application at sea. Apart from the frequency-range requirements, the necessary features include the following: (*a*) Satisfactory selectivity and high-frequency amplification. (*b*) No difficult adjustments, on account of the necessity for changing from one wave to another in a few seconds. (*c*) Ease of handling. These requirements naturally necessitate compromise. In general, high selectivity cannot be reconciled with ease of handling and quick wave-change, whilst very high amplification cannot be reconciled with stability over a wide range of frequencies. In addition, very high amplification is not of great practical value in ships because of the large amount of extraneous noise from running machines, aerial and stay vibrations, etc. The technical advantages of neutralization and its application to service receivers have been considered not only with regard to the receiver of Figs. 25 and 26 but also in connection with other models. Reference has been made in the paper to the fact that this receiver has had considerable application. It was designed several years ago to employ the naval R.7 valve which has an amplification factor of 7 and a grid-anode capacity of the order of 1 cm only. General-purpose valves of modern design have an amplification factor of the order of 14, and a grid-anode capacity of the order of 4 cm. Since feed-back is proportional (as a first approximation)to the product of amplification factor and grid-anode capacity, it follows that the need for neutralization is much less in the case of an R.7 valve than for one of a more recent type. Again, neutralization introduces at least one extra adjustment for the setting of a neutralizing condenser is, in general, different for different wave-ranges and may not even be constant over a single wave-range. Hence, if a neutralizing condenser is fitted, the stability of the receiver will be entirely in the hands of the operator. Failure to make this extra adjustment may lead to serious trouble, since when one receiver oscillates it may jam reception for other ships. The introduction of screen-grid valves is clearly a development for the later long-wave amplifiers, since these valves offer the possibility of high amplification without the trouble and difficulty of neutralization; later designs of long-wave receivers employ these valves.

With reference to Mr. Barton's remarks concerning the adoption of clockwise movement for inductance and capacities to correspond to increase of frequency, we think that Commander Daniel has stated the case for this convention as regards naval practice.

Mr. Lunnon has referred to the effects which occur when a fixed length of aerial is used with a short-wave band of frequencies. A conductor through a long trunk in a ship can be considered to be a transmission line leading from the transmitter to the aerial, and standing waves of current and voltage will be set up in it unless means are employed whereby the aerial and the transmitter-circuit impedances can be matched with the surge impedance of the trunk. Series or parallel connection of adjustable condensers at the coupling coil of the set constitutes one half of a suppression system and is certainly of value in getting more power into the aerial. These methods are employed in naval transmitters operating on narrow bands of frequency, but they are not used in the case of high-power short-wave transmitters working on the main aerial of a ship. Experience has shown that where the aerial system corresponds to a large number of $\frac{1}{4}$-wavelengths (as for main aerials), it is possible to put sufficient energy into it to obtain communication at the required ranges. We agree, however, that the employment of parallel or series condensers, possibly with additional coils at the coupling-coil, does increase the energy input to the aerial. In short-wave sets of lower power working through trunks to an aerial much shorter than the main aerial system, partial suppression has been employed (utilizing both the parallel and series systems of connections of variable condensers at the coupling coil of the set) when difficulty has been experienced in energizing the aerial. Where ship transmitters operate on a fixed wavelength, complete suppression of the standing waves has been employed. This, however, necessitates the insertion of a suitable tuning-circuit immediately adjacent to the main-deck insulator additional to the suppressing apparatus inserted at the coupling coil of the set. The naval use of "Mycalex" in standardized apparatus has not yet been sufficiently extensive to warrant a definite conclusion, but, so far as it has been employed in high-frequency circuits, the material has proved very satisfactory. Recent designs of adjustable high-voltage short-wave condensers, valve-holders, and coils, have employed "Mycalex" as insulation. The fact that the surface of "Mycalex" is not damaged by occasional arcing is a great factor in its favour. Our experience is that high-quality ebonite has proved on the whole more satisfactory than any other insulating material which, until recently, has been available for use with high-power coils under tropical conditions; we admit that it is far from being an ideal insulator. Aerial coils with ebonite combs have had years of use in ships under tropical conditions. It is not yet possible to say whether "Mycalex" will replace ebonite for use with high-voltage aerial coils.

In reply to Mr. Faulkner, the single overload relay system has proved adequate for the three transmitting valves employed in parallel in high-power ship transmitters. The envelopes of the silica valves can be seen from the gangway in front of the valve panels, and the glow resulting from a loss of vacuum is readily visible. In actual practice, however, valve failure from this

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cause is rare. The horn-gap fuses constitute the high-tension protection in case of any breakdown of the rectifier system, but it has also to be noted that the main circuit breakers in the primary d.c. supply to the main motor-alternators will also operate in the event of a severe overload. Fig. 11 shows that the anode key in the long-wave circuits is inserted between the transmitting-valve anodes and the anode blocking-condensers, the other terminal of which goes to the anode tapping-point on the aerial inductance. The valves used have fairly open grids so that this key cannot be employed for signalling unless suitable grid-potential control is also employed during spacing. This key opens the circuit during spacing, solely to ensure that there is no loss of received signal-strength from valve-damping across a portion of the aerial inductance when listening-through. The signalling-key of this transmitter is in the primary circuit of the a.c. supply and is the magnetic key shown in Fig. 10. The bobbin circuit for the anode key of Fig. 11 is placed across the front and middle contacts of the master transmitting-key so that the anode key always makes circuit before, and breaks after, the magnetic key for signalling. The d.c. control circuits are shown in Fig. 15. Interference troubles undoubtedly occur when the main ship transmitter is in use, and certain of the other aerials have to be earthed. Thus Fig. 17 indicates that the second office and the direction-finding office aerials can be earthed during transmission, but the transmitters of lower power can usually be employed simultaneously without serious interference to reception on any one of the lines. The reason for the adoption of separate grid-leaks for the particular transmitter circuit described in the paper was to avoid spurious oscillations which were found to occur when a common grid-leak was used. In this way the very high-frequency parasitic oscillation which occurred between the anodes and grids of the valves with a common grid-leak was avoided. The switching-off of the valve transmitting- and rectifying-filaments by excess current operating the overload relay has proved satisfactory and this system has been in use for over 10 years in various ship-transmitters. It should be observed that each of the rectifying valves takes a proportion of the H.T. supply voltage when the filaments are switched off. Thus the voltage stress is not across the transmitting valves alone. The combined long-wave and short-wave transmitter of Fig. 19 has been designed so that the valves are double-acting or in " push-pull " as regards the generation of high-frequency short-wave oscillatory current. The circuit arrangements are thus of symmetrical design and have proved stable with suitable neutralization of the valve anode-grid capacities. The chief difficulty in the design of this transmitter arose from the effects of secondary emission of the grids of the transmitting valves.

In answer to Mr. Turner's and Mr. Faulkner's remarks regarding the frequency-doubling circuit, Fig. 21, the circuits are neutralized by means of the adjustable condensers, so that there is no back-coupling. A H.T. supply to each stage is required and this is now indicated in the diagram. The overall efficiency of the combined circuits of Figs. 20 and 21 is less than that of the circuits of Fig. 20. However, when alternative transmissions on either frequency are desired the arrangements are useful and less costly than a second transmitter, while the extra space required is small. Transmissions can be made readily on either of the two wavelengths without tuning readjustments, so that operation is simple. A transmitter designed solely for the higher frequency corresponding to the doubler value would undoubtedly have a greater overall efficiency as regards H.T. power supply, but its initial cost would naturally be much greater than the cost of adding the doubler circuit to an existing transmitter. Again, other factors, such as the space available for transmitting apparatus, had to be taken into consideration when deciding whether a frequency-doubler circuit for an existing transmitter, or an additional transmitter, was to be employed.

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SOME MEASUREMENTS IN CORNWALL OF THE SIGNAL STRENGTH FROM 5XX.*

By J. H. Reyner, B.Sc., Associate Member.

(*Paper first received* 10*th October, and in final form* 19*th November*, 1929.)

Summary.

In connection with the testing of wireless receivers, information was desired concerning the field strength obtained from 5XX in Cornwall, this district being known to be poor from the point of view of wireless reception.

An investigation of the field strength was therefore carried out with the aid of portable equipment. This particular apparatus was simpler than that usually employed but appears from the consistency of the results to be reliable.

When the experiments were conducted, a curious effect was brought to light which appeared to be a radio shadow caused by the masts of the beam station at Bodmin. This gives rise to an interesting distribution of the contours in the Truro district.

The general order of field strength appears to coincide with theory up to this point, if due allowance is made for attenuation, but there is a rapid falling-off towards the extreme west of Cornwall.

(1) Apparatus Employed.

For certain reasons in connection with receiver testing it became desirable to obtain information regarding the signal strength of Daventry 5XX in Cornwall, as the evidence obtained from reception tests appeared to indicate a more rapid attenuation than normal. A simple portable equipment was therefore constructed, and a series of measurements was made covering a wide area. A small aerial and counterpoise system was used, the masts being in three sections so that they could readily be dismantled. The signals received on this aerial system were applied to a screened receiver comprising one stage of high-frequency amplification followed by an anode-bend rectifier.

A screened local oscillator was employed in conjunction with the equipment, the method of taking the measurements being as follows. The receiver was tuned in to 5XX and the rectified current in the anode circuit of the detector valve was measured by means of a sensitive microammeter. The receiver was then switched over to a dummy aerial having the same characteristics as the actual aerial system. In this circuit was included a coupling coil from the screened local oscillator, the output from which was adjusted to produce the same rectified current as before.

Provided that suitable obvious precautions are taken, such as ensuring that the oscillator is correctly tuned to the receiver, this method gives consistent and reliable results. The amplification of the receiver employed may vary from day to day, but since it is only used for the purpose of providing a visible indication of the rectified current the actual amplification is immaterial, the point which is important being the output from the oscillator.

* Reprinted from *Journal I.E.E.*, 1930, vol. 68, p. 181.

The oscillator was straightforward in character, consisting of a simple valve-maintained oscillating circuit to which a small pick-up coil was coupled. The inductance of this pick-up coil was kept small, only a few turns being employed, and the distance from the main tuning coil carrying the oscillating current was sufficiently great to render the capacity coupling negligible. Under such circumstances, the voltage induced in the pick-up coil is proportional to the current in the oscillating circuit and to the number of turns on the coil itself. The oscillator was so calibrated that the voltage induced in the pick-up coil was known in terms of the oscillating current, for each of the pick-up coils used.

A thermo-couple was included in the oscillating circuit, and the leads from this were taken to terminals on the case of the oscillator to which a meter could be connected externally. The H.T. voltage on the anode of the oscillating valve was made variable by means of a suitable high resistance, and by this system it was found practicable to control the oscillating current within wide limits and thereby to vary the induced voltage from 0·1 to 3 mV. This was found to be more than sufficient to cover the range of field strength encountered during the tests.

By means of this equipment, therefore, it is possible to measure, in any desired locality, the actual voltage in millivolts set up in the aerial by the radiations from 5XX. If the effective height of the aerial system is known it is possible to convert these figures into field-strength values. The best way of doing this is to measure the field strength from 5XX with the same apparatus at some point relatively close to Daventry such that the field strength is sufficiently large to be measured directly. For this purpose an absolute measurement of the signal strength from 5XX was taken at the author's laboratories at Elstree, and at the same time a reading was taken on the portable equipment. This enabled any readings obtained with the portable apparatus to be converted to millivolts per metre by direct comparison, and at the same time enabled the effective height of the aerial system to be ascertained. The dimensions of the aerial system are shown in Fig. 1.

For the absolute measurement of signal strength a large frame aerial of 4 ft. side was constructed and this was arranged to tune with a relatively small value of capacity to Daventry's wavelength. A sensitive valve voltmeter was connected across the tuned frame, and the reading on this meter was observed when the frame circuit was tuned to resonance.

The frame was then placed at right angles to the

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maximum position, in which case no signal was induced from 5XX, and a voltage was set up in the frame from a calibrated source. This voltage was adjusted until the deflection on the valve voltmeter was the same as before.

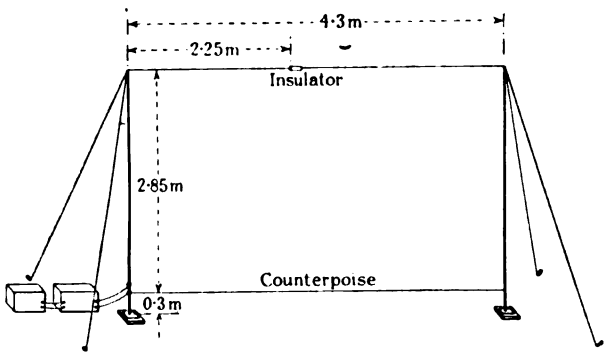

FIG. 1.—Arrangement of aerial system.

In order to check any serious error two methods were employed for inducing the necessary voltage in the frame aerial.

The first method consisted in the use of the same screened calibrated oscillator as was used with the portable aerial. Knowing the voltage induced in the various coupling coils per milliampere of oscillating current, it was a simple matter to determine the actual voltage induced in the frame. The second method consisted in introducing a 2-ohm resistance into the frame circuit. An oscillating current was passed through this resistance and the current was adjusted until the meter deflection was the same as before. The product of this current and the resistance gave the induced voltage.

Both methods gave consistent results in good agreement one with another. From the values of induced voltage so determined the field strength in millivolts per metre is easily evaluated from the well-known formulæ for frame aerial reception. As a result of these calculations, the field strength from 5XX at Elstree was found to be 12·5 millivolts per metre, a value which, as indicated in Section (3), is in good agreement with theory.

(2) Results Obtained.

The first tests consisted of a series of readings taken at selected points in the county in order to obtain some idea of the general contour to be expected. At first, there appeared to be little unusual, beyond the fact that the signal strength was lower than had been anticipated. Moreover, it was at once apparent that the contours were not circular or even approximately so in formation, being rather in the nature of a wedge in which the high signal-strength areas extended further west inland than on the coast. It was as if the coast-line were exercising some marked absorbing effect. Apart from this, however, the contours were not unusual, being of the general character indicated in Fig. 2.

Acting on this first rough contour, a number of further readings were taken and the results obtained were compared with the theoretical expectations. The figures were not in agreement and the discrepancies were for some time very puzzling. In some cases the signal strength would be distinctly lower than was anticipated, while in other cases it would be higher. Another curious effect was that in the Perranporth district the signal

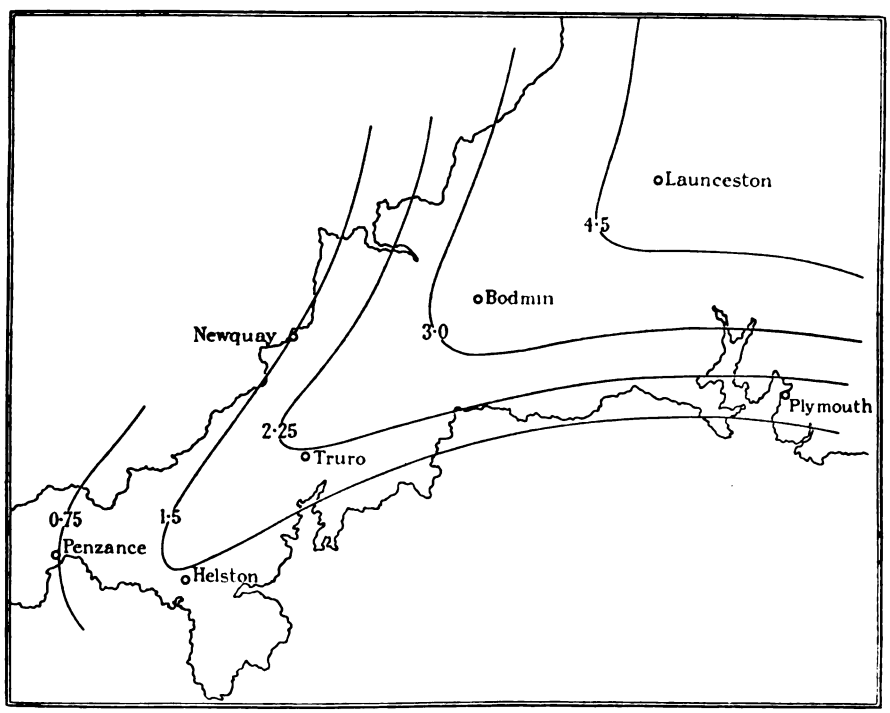

FIG. 2.—Apparent contours after first tests.

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strength on the coast-line is approximately half that on the hilltop, only two miles inland. This effect was confirmed by the local residents who had noticed a marked difference in actual reception in the two localities.

In the course of the investigations, however, the breasting of a hilltop one day suddenly brought into view the masts of the Bodmin beam station and this immediately suggested the possibility of a radio shadow. The results so far obtained were re-examined in the light of this possibility and appeared to give some confirmation of the view. Further investigations were therefore put in train in order to confirm or disprove the theory.

point such as Truro, the signal strength rises on either side and then falls away rapidly towards the coast-line. What is perhaps more peculiar is that as one proceeds farther west the signal strength rises at first instead of falling and then subsequently attenuates very rapidly.

Particular care was taken to confirm this point in order to determine whether the ridge of high field-strength was maintained completely round the shadow area or whether it would be observed at the two sides only. Definite evidence of good signal strength, however, was found at such places as Helston and Camborne, both of which points lie on the theoretical contour line

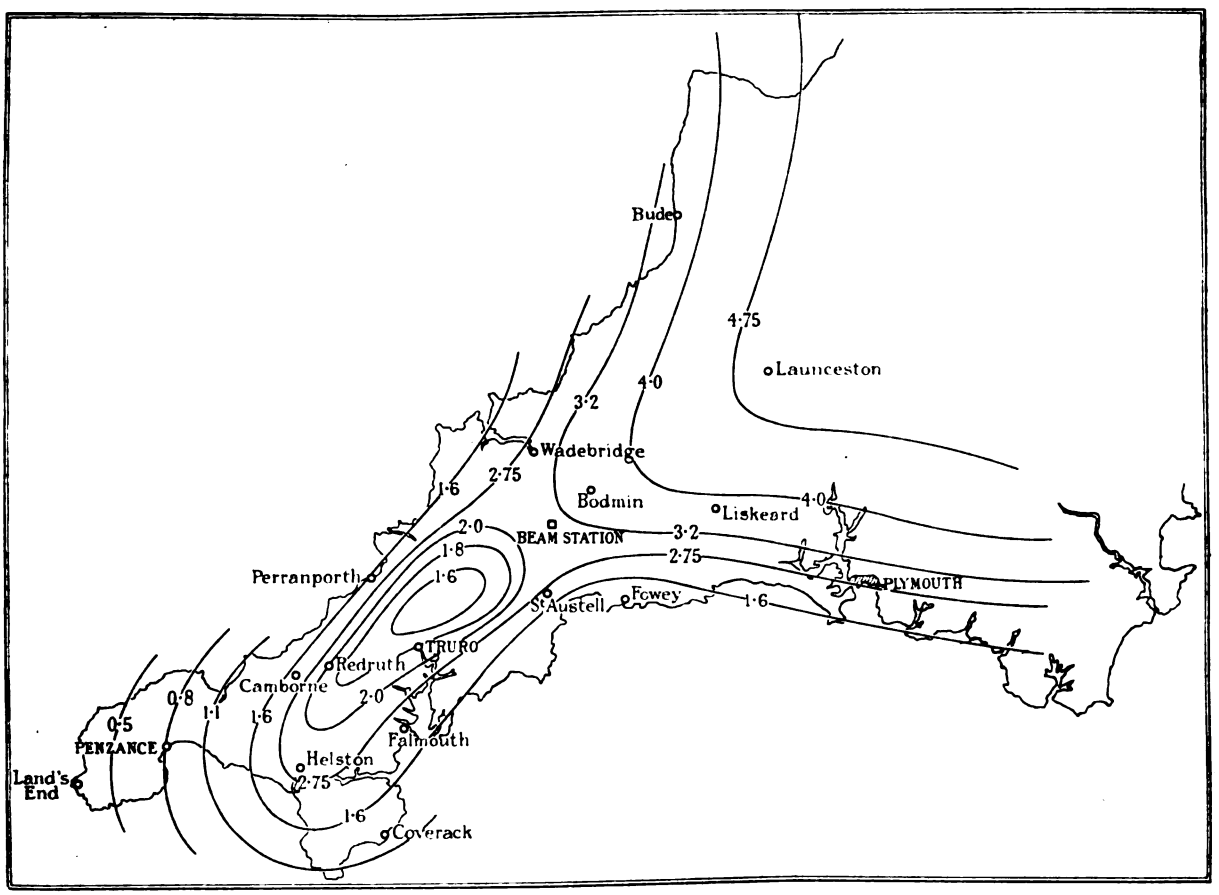

FIG. 3.—Final contours, showing shadow effect.

In every case the results obtained appeared to bear out the suggestion remarkably well, the final contours being shown in Fig. 3. On the north-east side of the Bodmin beam station the strength is of a normal order and in conformity with the contours shown in Fig. 2. At a point less than 10 miles to the south-west, however (on the other side of the Bodmin masts), the strength falls to approximately half its former value. Moreover, this region of bad reception extends in an almost direct line from Daventry through Bodmin as far as Redruth, a distance of nearly 25 miles.

A further interesting point is that surrounding this area of low signal strength is a ridge of good reception. Thus as one goes out towards the coast from a central

which one might expect. It appears, therefore, that this belt of good reception does actually exist immediately beyond the dead area around the Truro district.

The falling-off, once this high-field-strength ridge has been passed, is very rapid. In the short distance between Camborne and Penzance, for example, the signal strength falls to nearly one-third of its initial value, while the effects, already cited, of the very rapid drop in signal strength along the coast-line in such localities as Perranporth are also attributable to the same cause.

Whether the shadow effect is actually produced by the Bodmin masts is a point which can only be definitely confirmed by further experiments. There is, of course, the possibility that the curious configuration of the

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contours is due to mineral deposits. In order to obtain further information inquiries were made in various localities regarding any observed difference in the reception from 5XX after the erection of the Bodmin beam station. The station commenced operation some years ago and the information therefore is not reliable, but there is a distinct impression that the interference experienced during the reception of 5XX has been greater since the beam station was erected. Indeed, it is understood that representations were made to the Post Office to this effect when the station was first brought into commission, and investigations were made to see what interference was being produced by the beam station. No evidence of such interference was forthcoming and the matter was dismissed.

The suggested theory of a radio shadow, however, provides a satisfactory explanation. In certain districts the signal strength would fall considerably, in consequence of which the reception in those areas would be rendered more liable to interference from shipping and atmospherics, both of which are particularly troublesome in this part of England. The complaint of more interference, therefore, was apparently justified, but the nature of the interference was of a different character from what was anticipated.

(3) Analysis of Results.

No detailed analysis of the results has been made, having regard to present-day theories of wave propagation. Certain check calculations have been made, however, which appear to indicate that the results are of the order to be expected. All readings, of course, were taken during the daytime, the period being the month of June 1929, and no readings were taken after 7 p.m. as it was felt that there might be danger of interference by the indirect ray at later times than this.

The first check calculation was the effective height of the 5XX aerial as deduced from the observed field strength at Elstree. As the distance from Daventry is 90 km there is little likelihood of serious attenuation. An estimate of the attenuation was made from Sommerfeld's formula and a figure of 4 per cent taken.*

On this assumption, the un-attenuated field strength at Elstree would be 13·0 millivolts per metre, approximately. Thus the effective height of the 5XX aerial, based on an assumption of 60 amperes aerial current, would be 80·5 m.† The 5XX masts are 164 m high, the aerial being of T shape and the down-lead being approximately 130 m high. In the light of these figures, the value of effective height appears to be eminently reasonable.

* "The Attenuation of Wireless Waves due to the Resistance of the Earth," *Journal I.E.E.*, 1926, vol. 64, p. 766.

† The wavelength of 5XX at the time of the tests was 1 562·5 m.

The second check was on the signal strength actually received in Cornwall as compared with the theoretical value calculated from the usual formula. In view of the distorted contours it is somewhat difficult to decide what shall be taken as the mean field strength at any given distance from Daventry. The wedge-shaped formation is also troublesome, for there are two ways of regarding this. One is to consider the strength at the peak as correct and to assume that the falling-away at the sides is due to some absorption effect produced at the coast-line. It is not quite clear how such an effect could arise. The second way of regarding the contours is to assume that the true contours are at the coast-line and that the extension of the contour towards the west inland is due to the Bodmin masts in a similar but opposite manner to the shadow effect on the other side of Bodmin. There appears to be some justification for this point of view because the attenuation deduced on the first assumption is very small, rather less, indeed, than that over sea water.

Probably the most satisfactory procedure is to take a mean value and this has been done in Table 1.

Table 1.

Comparison of Actual and Theoretical Field Strength.

Distance from 5XX	Theoretical field strength (un-attenuated)	Mean observed field strength	Ratio
km	mV/m	mV/m	per cent
280	4·18	3·8	90·5
305	3·81	3·2	84·0
335	3·5	2·5	71·5
360	3·27	1·6	49·0
385	3·04	0·8	26·3

It will be observed that up to 300 km the attenuation is fairly small but not unreasonably so, and the figures coincide approximately with the values obtained from Sommerfeld's theory. Beyond this point, however, the attenuation is increasingly rapid, and at Land's End, which is only 400 km away, the field strength is less than one-quarter of its theoretical value.

As was previously stated, this analysis is very rough and was intended merely as a check on the reasonableness of the figures obtained. The author has not had time nor is he sufficiently well acquainted with the formulæ to make a comprehensive analysis, but it is hoped that the results quoted will be of assistance to more qualified investigators in their endeavours to define the laws governing the propagation of wireless waves.

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THE VALVE-MAINTAINED QUARTZ OSCILLATOR.*

By J. E. P. VIGOUREUX, M.Sc., Graduate.

[From the National Physical Laboratory : communicated by permission of the Radio Research Board.]

(*Paper first received 24th June, and in final form 18th October*, 1929.)

SUMMARY.

A quartz resonator connected between the grid of a valve and either of the other electrodes gives rise to oscillations in the plate circuit when the constants of that circuit satisfy certain conditions. The frequency of the oscillations depends mainly upon the natural frequency of the quartz, and is, therefore, fairly constant when the temperature is constant. In addition to temperature effects, which are not considered in the present paper, variations of frequency can be produced by (i) variations of the air-gaps of the resonator, i.e. the air-gaps between the quartz and the two electrodes of the resonator; (ii) variations of the constants of the plate circuit; and (iii) variations of the inter-electrode capacities and conductances. The object of the investigation is the study of these variations, both theoretically and experimentally.

In the theoretical treatment, the quartz is replaced by a circuit made up of an inductance, a resistance and a capacity in series, the whole being in parallel with a second capacity. The validity of such a substitution has been studied experimentally by Dye.

The combination of quartz resonator, valve and plate circuit is represented by a network of admittances, from which three circuit equations are obtained; to these is added the fundamental equation of the valve. These four equations yield a single linear differential equation of the fifth order, with constant coefficients. In the case under investigation, it is only required to find the frequency and the amplitude of oscillations after the latter have reached a constant amplitude. This consideration affords a means of reducing the differential equation to one of the same type, but of the second order. From the latter equation, of which the general solution is known, formulæ are developed for the changes of frequency resulting from variations in the resistance and capacity of the plate circuit, in the resistance of the grid leak, in the inter-electrode capacities, and in the air-gap. These formulæ provide useful information as to the influence of each variable on the frequency, and can with advantage be applied to the choice of the best circuits and the best valve for a quartz oscillator.

As the oscillations build up after switching on, the average plate conductance of the valve alters gradually. When the amplitude of oscillation has reached its final value, the value of the average plate conductance is such as to annul the damping term, i.e. the coefficient of dy/dt in the differential equation. From this condition the average plate conductance can be determined for any values of the circuit variables. The amplitude of oscillations in the plate circuit is a function of this average plate conductance.

Formulæ are developed which give the conditions for the maintenance of oscillations, and an investigation is made of the variations of amplitude with variations in (*a*) capacity in the plate circuit, (*b*) resistance in the plate circuit, and (*c*) grid conductance.

Points of importance in the choice of circuits and valves are deduced from the formulæ and the theoretical curves

Two types of quartz oscillator are investigated, namely, the cases in which the quartz resonator is connected between (i) the grid and the plate of the valve, and (ii) the grid and the filament of the valve.

A fairly complete experimental investigation is described. The admittances of the valve and of the circuits were measured, and the equivalent electrical network of the quartz was determined. The behaviour of the quartz oscillator was then investigated by altering in turn each variable in the circuits. The theoretical and experimental results are compared and discussed.

The paper does not deal with circuits especially devised for obtaining large outputs from quartz crystals. Its object is the study of the influence of the circuit variables on the frequency and on the power generated as regards the simple Pierce circuits.

INTRODUCTION.

W. G. Cady† and G. W. Pierce‡ have shown that quartz resonators can be used for maintaining oscillations in a valve circuit. The frequency of the oscillations depends almost entirely on the natural frequency of vibration of the quartz, and is, therefore, very constant. The combination of resonator and valve circuit is known as a "quartz oscillator." Quartz oscillators are at present widely employed for controlling the frequency of wireless transmitting stations, in the same way that an ordinary master oscillator was used formerly, and give excellent results. They are also very convenient for calibrating wavemeters, and are extensively used for that purpose in America.

The circuits considered in this paper are all based on Pierce's circuit. A single valve is used, with the quartz connected between the grid and the plate, or between the grid and the filament, and, under certain conditions, oscillations are maintained in the plate-filament circuit. This latter circuit may be a high non-inductive resistance, of approximately the same value as the valve impedance; but for powerful oscillations an oscillatory circuit is used, made up of a condenser and an inductance coil in parallel, and oscillations are maintained for a wide range of condenser settings.

The type of connection in which the resonator is connected between the plate and the grid of the valve will be called the "plate-to-grid connection," and the other type the "grid-to-filament connection."

Besides temperature effects, which need not be considered here, variations in frequency are caused by (*a*) variations of the air-gap; (*b*) variations of the constants of the plate circuit; and (*c*) variations of the inter-electrode capacities and conductances.

* Reprinted from *Journal I.E.E.*, 1930, vol. 68, p. 265.

† See Reference (1).

‡ *Ibid.* (2).

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E. M. Terry* and J. W. Wright† have recently given a method of predicting the variations of frequency due to variations of the circuit constants. We are dealing here with the same problem and, in addition, an attempt is made to obtain an idea of the power to be expected for various values of the circuit constants.

(1) THEORETICAL TREATMENT.

The Fundamental Equations.

The circuit may be represented in a general way by Fig. 1, in which G_1, G_2 and G_3 represent admittances. They include the inter-electrode admittances as well as the admittances of external circuits.

Let g be the anode conductance of the valve and m its amplification factor. The fundamental equation of the valve is

$$i_1 + i_2 = g(v + me) \quad . \quad . \quad . \quad . \quad (1)$$

with the currents and potential differences indicated in Fig. 1.

It is assumed that the variations of current are such as would be produced by a straight-line characteristic

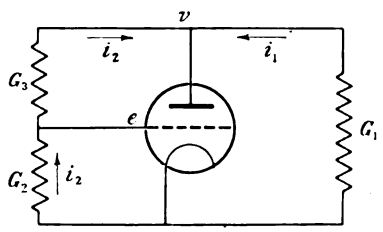

Fig. 1.

for the valve. This assumption is never realized in practice, for, once oscillations start, they must of necessity increase in amplitude until the anode conductance g alters in such a way that the exponential term in the solution of the differential equation connecting current and time becomes unity. Here, however, it is assumed that g does not change. It is also assumed that the conductance between the grid and the filament is constant throughout a complete oscillation. This is not, in general, the case, because for part of the cycle the grid is at a negative potential with respect to the filament, when the conductance is only that of the grid leak; and for the rest of the cycle it is at a positive potential with respect to the filament, and the conductance becomes larger owing to thermionic effect.

In addition to equation (1) we can write down the three circuit equations:—

$$\left.\begin{aligned} i_1 &= -G_1 v \\ i_2 &= -G_2 e \\ i_2 &= -G_3(v - e) \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad (2)$$

In general, G_1, G_2 and G_3 will all contain the differential operator D and the resultant equations in i_1, i_2, v or e will be identical, which means that the complementary function is of the same form for i_1, i_2, v and e.

From equations (1) and (2) we obtain

$$\begin{vmatrix} 1 & 1 & -g & -mg \\ 1 & 0 & G_1 & 0 \\ 0 & 1 & 0 & G_2 \\ 0 & 1 & G_3 & -G_3 \end{vmatrix} y = 0$$

where y stands for either i_1, i_2, v or e.

This determinant gives the simple equation

$$\left[G_3\{G_1 + G_2 + (m+1)g\} + G_2(G_1 + g)\right]y = 0 \quad . \quad (3)$$

Expressions must now be obtained for G_1, G_2 and G_3 in terms of the circuit constants and of the valve constants.

Consider G_1 first; it is a circuit similar to that in Fig. 2, in which the capacity C includes the capacity C_1

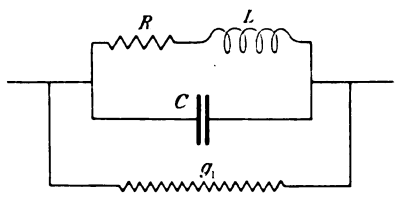

Fig. 2.

of the plate to the filament. The capacity C is therefore always greater than zero. The inter-electrode plate-to-filament conductance is denoted by g_1.

The conductance of a capacity C is denoted by DC, and that of an inductance L by $1/DL$, where D stands for the differential operator d/dt.

We therefore have

$$G_1 = \{1/(R + DL)\} + DC + g_1 \quad . \quad . \quad (4)$$

We also have to evaluate G_2 and G_3. Let g_2 and C_2 be the inter-electrode conductance and capacity between grid and filament, and let g_4 be the conductance of a

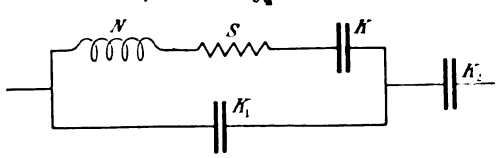

Fig. 3.

grid leak. That part of G_2 due to the above quantities is $(g_2 + g_4 + DC_2)$.

Let g_3 and C_3 be the inter-electrode conductance and capacity between the plate and the grid. That part of G_3 due to them is $(g_3 + DC_3)$.

The admittance of the quartz circuit is added to either of these, depending upon whether the quartz is placed between the plate and grid, or between the filament and grid.

It has been shown by W. G. Cady,* D. W. Dye† and others, that a fairly accurate electrical representation of the behaviour of the quartz is given by the circuit NSKK$_1$ of Fig. 3; the capacity K_1 is approximately equal‡ to the static capacity between the two faces of the quartz perpendicular to the electrical axis e, whilst K is only an equivalent capacity depending upon the

* See Reference (3). † *Ibid.*, (9).

* See Reference (1). † *Ibid.*, (4). ‡ *Ibid.*, (5).

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piezo-electric and elastic constants. The resistance S accounts for the damping of the quartz vibrator, K_2 is the capacity of the air-gap between the electrodes, and K_3 the stray capacity between the electrodes and between the leads.

The admittance of the quartz will then be

$$DK_3 + 1\Big/\left[\frac{1}{DK_2} + 1\Big/\left\{DK_1 + 1\Big/\left(DN + S + \frac{1}{DK}\right)\right\}\right]$$

$$= DK_3 + \frac{DK_2(D^2NKK_1 + DSKK_1 + K_1 + K)}{D^2NK(K_1+K_2) + DSK(K_1+K_2) + K_1 + K_2 + K}$$

Using Dye's notation, we obtain

$$K_1 + K_2 = K_t \quad . \quad . \quad . \quad . \quad . \quad (5)$$

Then the admittance of the quartz system is

$$D\left[K_3 + \frac{D^2NK + DSK + 1 + (K/K_1)}{D^2NK + DSK + 1 + (K/K_t)} \cdot \frac{K_1K_2}{K_t}\right]$$

For a further simplification, let

$$D^2NK + DSK + 1 + \frac{K}{K_1} = \sigma \quad . \quad . \quad (6)$$

$$D^2NK + DSK + 1 + \frac{K}{K_t} = \nu \quad . \quad . \quad (7)$$

The admittance of the quartz is then given by

$$DK_3 + D(K_1K_2/K_t)(\sigma/\nu)$$

The values of G_1, G_2 and G_3 are thus obtained in terms of the circuit constants and of the valve constants. It remains only to substitute these values in equation (3), and to use this equation to find the condition for maintenance of oscillations, the frequency, and the amplitude of oscillations.

Before dealing with the complicated equation resulting from the substitution of G_1, G_2 and G_3 in equation (3), consider the simple general equation of an oscillating system:

$$(D^2A + DB + C)y = 0$$

The solution is obtained by solving the quadratic

$$\lambda^2A + \lambda B + C = 0$$

which gives

$$\lambda = \left\{-B \pm \sqrt{(B^2 - 4AC)}\right\}/2A$$

and the complementary function is

$$y = F\epsilon^{\left\{-\frac{B}{2A} + \sqrt{\left(\frac{B^2}{4A^2} - \frac{C}{A}\right)}\right\}t} + F'\epsilon^{\left\{-\frac{B}{2A} - \sqrt{\left(\frac{B^2}{4A^2} - \frac{C}{A}\right)}\right\}t}$$

where F and F' are constants.

If $B^2 > 4AC$ the function increases or decreases exponentially, but there are no oscillations.

If $B^2 < 4AC$, the above solution is equivalent to

$$y = \epsilon^{-\frac{B}{2A}t}\left\{F'' \cos\sqrt{\left(\frac{C}{A} - \frac{B^2}{4A^2}\right)}t + F''' \sin\sqrt{\left(\frac{C}{A} - \frac{B^2}{4A^2}\right)}t\right\}$$

and oscillations take place. These oscillations may be either damped or maintained.

If $B/2A > 0$ they will be damped, and any disturbance set up will gradually die away; if, on the other hand, $B/2A < 0$ oscillations are maintained, and, starting from any small disturbance, their amplitude increases until the original quantity $B/2A$ is balanced by the increase in damping which in general accompanies an increase in amplitude. At this moment the effective B is zero, and the equation of oscillations is

$$y = F'' \cos\sqrt{\left(\frac{C}{A}\right)}t + F''' \sin\sqrt{\left(\frac{C}{A}\right)}t$$

The circular frequency ω is thus given by the equation

$$\omega^2 = C/A$$

The final amplitude of y must be some function of the original value of $-B/2A$; it is not easy to estimate this function, in general, but in some cases the initial value of $-B/2A$ may be used to predict the final amplitude.

The following rules for the study of oscillations can thus be stated:

(i) If the coefficients of D^2 and D^0 are opposite in sign, no oscillations can take place.

(ii) If the coefficients of D^2 and D are of the same sign, no oscillations can be maintained.

(iii) If the coefficients of D^2 and D^0 are of the same sign, and opposite in sign to that of D, and if $4AC$ is greater than B^2, oscillations will be maintained at a frequency eventually given by $(1/2\pi)\sqrt{(C/A)}$.

(iv) The original value of $-B/2A$ may, in some cases, afford a means of calculating the final amplitude.

Plate-to-Grid Connection.

Returning now to the particular problem of the quartz oscillator, suppose that the quartz resonator is connected between the grid and the plate of the valve. In this case

$$G_1 = \left\{1/(DL + R)\right\} + DC + g_1 \quad . \quad . \quad . \quad . \quad (8)$$

$$G_2 = g_2 + g_4 + DC_2 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (9)$$

$$G_3 = g_3 + D(C_3 + K_3) + D(K_1K_2/K_t)(\sigma/\nu) \quad . \quad (10)$$

The conductance of the grid leak is denoted by g_4. As for g_2 and g_3, they include not only the direct-current conductance of the glass and valve-holder, which is usually extremely small, but also the effective conductance due to dielectric loss in the glass and holder between the electrodes and the thermionic effect, if any. The direct-current component and the dielectric component of g_3 are different in their effects, which will be discussed later in the paper.

Substituting for G_1, G_2 and G_3 in equation (3), the following equation is obtained:—

$$\left[D(K_1K_2/K_t)(\sigma/\nu) + g_3 + D(C_3+K_3)\right]$$
$$\left[g_1 + g_2 + g_4 + (m+1)g + D(C+C_2) + 1/(DL+R)\right]$$
$$+ \left[g_2 + g_4 + DC_2\right]$$
$$\left[g_1 + g + DC + 1/(DL+R)\right] = 0 \quad . \quad . \quad (11)$$

The quantity operated on is y, standing for i_1, i_2, v or e, and will not be repeated each time.

For brevity, we can write

$$C_2 + C_3 + K_3 = C_5 \quad . \quad . \quad . \quad . \quad (12)$$

Equation (11) can be put in the form

$$(K_1K_2/K_t)(\sigma/\nu)\alpha + C_5 = 0 \quad . \quad . \quad (13)$$

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in which α is a fraction whose numerator and denominator are both of the third degree in D.

The numerator of α will be

$$D^3L(C+C_2) + D^2\left[R(C+C_2)+\left\{g_1+g_2+g_4+(m+1)g\right\}L\right] + D\left[1+\left\{g_1+g_2+g_4+(m+1)g\right\}R\right] \quad . \quad (14)$$

and the denominator will be a similar expression, with a term independent of D.

The evaluation of α is greatly simplified by neglecting mRg in comparison with unity. This can be done only when G_1 is an oscillatory circuit. The case when it is made up of a high resistance should be treated separately, and is simpler.

If the frequency were, say, 100 000 cycles per sec., and a low-impedance valve were used, the quantities considered might be as follows:

$L = 2\,000$ microhenrys
$R = 10$ ohms
$m = 10$
$g = 0{\cdot}0003$ mho

Then $mRg = 0{\cdot}03$.

This is, however, an extreme case. It will be proved later that the resistance R of the coil L should be as small as possible if large powers are to be obtained. Moreover, quartz oscillators are in general worked at frequencies well above 100 000 cycles per sec. The value of mRg is, therefore, in most cases well under $0{\cdot}01$ and can be neglected in comparison with unity.

From equations (6), (7) and (13) we obtain

$$\left\{D^2NK + DSK + 1 + (K/K_1)\right\}(K_1K_2/K_t)\alpha + (D^2NK + DSK + 1 + K/K_t)C_5 = 0$$

$$\therefore\ D^2NK + DSK + 1 + \frac{K}{K_t}\cdot\frac{\alpha K_2 + C_5}{\alpha(K_1K_2/K_t) + C_5} = 0 \quad (15)$$

For brevity, we can write

$$1 + \frac{K}{K_t}\frac{\alpha K_2 + C_5}{\alpha(K_1K_2/K_t) + C_5} = \beta \quad . \quad . \quad (16)$$

The equation of oscillations is then

$$D^2NK + DSK + \beta = 0 \quad . \quad . \quad . \quad (17)$$

and β can be put into the form

$$\frac{\phi_1(D^2) + D\phi_2(D^2)}{\phi_3(D^2)}$$

If it were possible to obtain an approximation for D^2, the equation would take the simple form

$$D^2NK + DSK + D\beta_2 + \beta_1 = 0 \quad . \quad . \quad (18)$$

where β_1 and β_2 are real functions of the variables of the circuits.

When oscillations are just on the point of starting, or when oscillations are being maintained at a constant amplitude, the damping term is zero, that is

$$\beta_2 + SK = 0 \quad . \quad . \quad . \quad . \quad (19)$$

Therefore, also,

$$D^2NK + \beta_1 = 0 \quad . \quad . \quad . \quad . \quad (20)$$

The latter equation gives the required approximation for D^2.

It is here convenient to define a frequency f_0 by the formulæ

$$\left.\begin{aligned}\omega_0 &= 2\pi f_0\\ \omega_0^2NK &= 1\end{aligned}\right\} \quad . \quad . \quad . \quad . \quad (21)$$

The frequency f_0 is that frequency at which the quartz vibrates when it is used as a resonator with no air-gap.

Having thus defined ω_0, let C_0 be a capacity determined by the relation

$$\omega_0^2LC_0 = 1 \quad . \quad . \quad . \quad . \quad . \quad (22)$$

The capacity C_0 will be that value of C, including the inter-electrode plate-to-filament capacity C_1, for which the natural frequency of the oscillatory circuit is equal to the natural frequency of vibration of the piece of quartz, used as a resonator with no air-gap.

Having written $\beta = \beta_1 + D\beta_2$

we can write $\alpha = \alpha_1 + D\alpha_2$

where α_1 and α_2 are also real functions of the circuit variables.

Equation (20) gives

$$D^2 = -\omega_0^2\beta_1 \quad . \quad . \quad . \quad . \quad . \quad (23)$$

Substituting this in the expression for α as derived from equation (11), we find that the numerator of α is $G + DH$, where

$$G = C + C_2 - (C_0/\beta_1) \quad . \quad . \quad . \quad (24)$$

and

$$H = -(C_0/\beta_1)\left[R(C+C_2)+\left\{g_1+g_2+g_4+(m+1)g\right\}L\right] \quad (25)$$

If we write

$$U = C - (C_0/\beta_1) \quad . \quad . \quad . \quad . \quad (26)$$

and

$$W = -(C_0/\beta_1)\left[RC + (g_1+g)L\right] \quad . \quad . \quad (27)$$

we find that the denominator of α is $E + DF$, where

$$E = \frac{C_3 + K_3}{C_5}G + \frac{C_2}{C_5}U + g_3\frac{H}{C_5} + (g_2+g_4)\frac{W}{C_5}$$

and

$$F = \frac{C_3 + K_3}{C_5}H + \frac{C_2}{C_5}W - \frac{1}{\omega_0^2\beta_1}g_3\frac{G}{C_5} - \frac{1}{\omega_0^2\beta_1}(g_2+g_4)\frac{U}{C_5}$$

Finally, since

$$\alpha = \frac{G + DH}{E + DF} \quad . \quad . \quad . \quad . \quad . \quad (28)$$

we have

$$\alpha_1 = \frac{GE + \omega_0^2HF\beta_1}{E^2 + \omega_0^2F^2\beta_1} \quad . \quad . \quad . \quad (29)$$

and

$$\alpha_2 = \frac{HE - GF}{E^2 + \omega_0^2F^2\beta_1}$$

$$HE - GF = -\frac{C_2}{C_5}\frac{mg}{\omega_0^2\beta_1}\left[C - \frac{C_0}{\beta_1} - \frac{C_2}{m}\left(1 + \frac{R}{g}\frac{C_0}{L}\frac{1}{\beta_1}\right)\right] + \frac{1}{\omega_0^2\beta_1C_5}\left[g_1C_2^2 + g_3(\omega_0^2H^2\beta_1 + G^2) + (g_2+g_4)(\omega_0^2WH\beta_1 + U^2)\right] \quad . \quad (30)$$

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The expressions for E and F are

$$E = C + C_2 - \frac{C_2^2}{C_5} - \frac{C_0}{\beta_1} + g_3\frac{H}{C_5} + (g_2 + g_4)\frac{W}{C_5} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (31)$$

$$F = -\frac{C_0}{\beta_1}\left[R\left(C + C_2 - \frac{C_2^2}{C_5}\right) + \left(m\frac{C_3 + K_3}{C_5} + 1\right)gL + g_1L + \frac{C + C_2 - (C_0/\beta_1)}{C_5}g_3L + \frac{C + C_3 + K_3 - (C_0/\beta_1)}{C_5}(g_2 + g_4)L\right] \quad . \quad (32)$$

The expressions for β_1 and β_2 in terms of α_1 and α_2 can be obtained from equation (16). They are

$$\beta_1 = 1 + \frac{K}{K_t}\,\frac{(\alpha_1K_2 + C_5)\{\alpha_1(K_1K_2/K_t) + C_5\} + \alpha_2^2\omega_0^2\beta_1(K_1K_2^2/K_t)}{\{\alpha_1(K_1K_2/K_t) + C_5\}^2 + \alpha_2^2\omega_0^2\beta_1(K_1K_2/K_t)^2} \quad . \quad . \quad . \quad . \quad . \quad (33)$$

$$\beta_2 = \frac{\alpha_2C_5K(K_2/K_t)^2}{\{\alpha_1(K_1K_2/K_t) + C_5\}^2 + \alpha_2^2\omega_0^2\beta_1(K_1K_2/K_t)^2} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (34)$$

When oscillations are just on the point of starting, or are being maintained at a constant amplitude, $\beta_2 + SK = 0$, and therefore β_2 is negative, and of the same order of magnitude as SK. Equation (34) shows that α_2 must be negative when β_2 is negative.

When $\alpha_2 < 0$

$$GF > HE$$

$$G > \frac{HE}{F}$$

Hence $$\alpha_1 = \frac{GE + \omega_0^2HF}{E^2 + \omega_0^2F^2} > \frac{H}{F}$$

Moreover, since $\alpha_2 < 0$, equation (30) shows that $C > C_0/\beta_1$.

Under these conditions, it is easily seen from equations (25) and (32) that H/F is positive, that is, greater than zero; hence α_1 is also greater than zero, and $\alpha_1(K_1K_2/K_t) + C_5$ is of the same order of magnitude as C_5.

Equation (34) gives two negative roots for α_2 in terms of β_2. Since K/K_t is small, β_1 does not differ greatly from $1 + (K/K_t)$, that is, from unity, whatever be the value of α_2. The order of magnitude of one root of α_2 is therefore small and equal to SC_5. The second root is large, and its order of magnitude is $1/(\omega_0^2SK_1)$. This second root must, however, be rejected, for equations (29) and (30) show that the numerical value of α_2 is, except in some cases, over a very small range of values of C, less than $1/(\omega_0^2SK_1)$.

Since α_2 is small, we may write

$$\beta_1 = 1 + \frac{K}{K_t}\cdot\frac{\alpha_1K_2 + C_5}{\alpha_1(K_1K_2/K_t) + C_5} \quad . \quad . \quad (35)$$

and $$\beta_2 = \frac{\alpha_2C_5K}{\{\alpha_1(K_1K_2/K_t) + C_5\}^2}\cdot\frac{K_2^2}{K_t^2} \quad . \quad . \quad (36)$$

If the capacity C be of such a value that $\beta_2 + SK < 0$, oscillations start and build up. As the amplitude of the oscillations increases, the modulus $|\beta_2 + SK|$ decreases, and therefore $|\beta_2|$ becomes more and more nearly equal to SK. The change in β_2 can be produced only by a change in the anode conductance g. Finally the amplitude of oscillations reaches such a value that the corresponding mean value of g makes the expression $\beta_2 + SK$ equal to zero. The amplitude ceases to increase, and the oscillations are maintained at constant amplitude.

The frequency of the oscillations depends upon β_1, and therefore upon α_1. This quantity changes with g, so that the frequency of the oscillations changes slightly during the process of building up. The time taken for this building up is, in general, of the order of 1 or 2 seconds.

The frequency with which we are concerned is, of course, the frequency corresponding to the constant amplitude. When the amplitude has reached its constant value, $\beta_2 + SK = 0$ and the order of magnitude of α_2 is SC_5. This being very small, and the denominator $E^2 + \omega_0^2F^2$ of α_2 being very small, it is necessary that the numerator $HE - GF$ should be very small, namely, of the order of $SC_5C_0^2$. When the amplitude has reached a constant value, it is, therefore, permissible to write

$$\alpha_1 = \frac{G}{E} = \frac{H}{F} = \frac{C + C_2 - (C_0/\beta_1)}{C + C_2 - \frac{C_2^2}{C_5} - \frac{C_0}{\beta_1} - (g_2 + g_4)\left[\frac{C_0}{C_5\beta_1}(RC + gL)\right] - g_3\left[\frac{C_0}{C_5\beta_1}\{R(C + C_2) + (m + 1)gL\}\right]} \quad . \quad (37)$$

This approximation will give a reasonably accurate value for the frequency. The effect of g_2, g_3, g_4 and R on the frequency will be considered later. It will be sufficient here to neglect the terms in g_2, g_3 and g_4, and to write

$$\alpha_1 = \frac{C + C_2 - C_0/\beta_1}{C + C_2 - (C_2^2/C_5) - (C_0/\beta_1)} \quad . \quad (38)$$

For brevity, we shall write α and β instead of α_1 and β_1.

The frequency of oscillations is given by

$$\omega^2 = \omega_0^2\beta$$

and, since β never departs much from unity,

$$f = f_0(\tfrac{1}{2} + \tfrac{1}{2}\beta) \quad . \quad . \quad . \quad . \quad . \quad . \quad (39)$$

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The ratio of the coefficient of D to that of D^2 is

$$S K\omega_0^2 + \beta_2\omega_0^2 = S K\omega_0^2 + K\left(\frac{K_2}{K_t}\right)^2 \frac{g_1C_2^2 + g_3(\omega_0^2H^2 + G^2) + (g_2 + g_4)(\omega_0^2WH + U^2)}{E^2\left\{1 + (\omega_0^2F^2/E^2)\right\}\left\{\alpha(K_1K_2/K_t) + C_5\right\}^2}$$

$$- \frac{K(K_2/K_t)^2 mgC_2\left[C - \frac{C_0}{\beta} - \frac{C_2}{m}\left(1 + \frac{R}{g}\frac{C_0}{L}\right)\right]}{E^2\left\{1 + (\omega_0^2F^2/E^2)\right\}\left\{\alpha(K_1K_2/K_t) + C_5\right\}^2} \quad . \quad . \quad . \quad (40)$$

This expression shows that S, g_1, g_2, g_3 and g_4 all tend to oppose the production of oscillations. It is also permissible to state that they tend to make the final amplitude smaller. In the case of the grid leak g_4 this rule, however, is not rigid, because the grid leak may alter the grid potential so as to cause oscillations to take place about a more favourable point on the characteristic curve of the valve.

If we substitute for α by equation (38), remembering that this is only an approximation which assumes $HE = GF$, we find for the ratio of the terms in D and D^2:—

$$S K\omega_0^2 + K\left(\frac{K_2}{K_t}\right)^2 \frac{g_1C^2 + g_3(\omega_0^2H^2 + G^2) + (g_2 + g_4)(\omega_0^2WH + U^2)}{\left(1 + \frac{\omega_0^2F^2}{E^2}\right)\left(\frac{K_1K_2}{K_t} + C_5\right)^2\left(C + C_2 - \frac{C_0}{\beta} - \frac{C_2^2}{(K_1K_2/K_t) + C_5}\right)^2}$$

$$- K\left(\frac{K_2}{K_t}\right)^2 \frac{mgC_2\left[C - \frac{C_0}{\beta} - \frac{C_2}{m}\left(1 + \frac{R}{g}\frac{C_0}{L}\right)\right]}{\left(1 + \frac{\omega_0^2F^2}{E^2}\right)\left(\frac{K_1K_2}{K_t} + C_5\right)^2\left(C + C_2 - \frac{C_0}{\beta} - \frac{C_2^2}{(K_1K_2/K_t) + C_5}\right)^2} \quad . \quad (41)$$

When C is gradually increased from zero, the expression is positive and there can be no oscillations. When C reaches the value

$$C = \frac{C_0}{\beta} + \frac{C_2}{m}\left(1 + \frac{R}{g}\frac{C_0}{L}\right)$$

the expression is still positive; a small increase in C, however, makes it zero; a further increase makes it negative, and oscillations start. As C is still further increased, the expression quickly reaches a minimum, then increases again, gradually, until a certain large value of C is attained for which it becomes zero. At this point the oscillations stop. It is easy to find this limit, because for large values of C, G and F are large compared with ω_0H and ω_0F. It is given by

$$(mgC_2/C - C_0) = g_2 + g_3 + g_4 + S\omega_0^2\left\{K_1 + C_5(K_t/K_2)\right\}^2 \quad (42)$$

The lower limit of C is given by

$$C > \frac{C_0}{\beta} + \frac{C_2}{m}\left(1 + \frac{R}{g}\frac{C_0}{L}\right) \quad . \quad . \quad . \quad (43)$$

Closer approximations to this lower limit, or "cut-off point," will be obtained in the Sections dealing with amplitude, pp. 49 and 50. Equation (41) will also be used to study the changes in amplitude with respect to C and to the other circuit variables.

Relation between frequency and total capacity in the oscillatory circuit.—It is convenient to work in frequency differences above f_0. The difference

$$\Delta f = f - f_0 = \tfrac{1}{2}f_0(\beta - 1)$$

$$\therefore \frac{\Delta f}{f_0} = \frac{1}{2}\frac{K}{K_t}\frac{\alpha K_2 + C_5}{\alpha(K_1K_2/K_t) + C_5} \quad . \quad . \quad (44)$$

It has been shown (p. 45) that $\alpha > 0$. Over the range

$$0 < \alpha < \infty$$

we have

$$\tfrac{1}{2}(K/K_t) < (\Delta f/f_0) < \tfrac{1}{2}(K/K_1)$$

Hence $(\Delta f/f_0)$ is always smaller than 1/280, which is the theoretical value of $\frac{1}{2}(K/K_1)$ for good quartz,* and is rarely reached in practice with resonators vibrating along the electric axes.

We have

$$\alpha = \frac{C + C_2 - (C_0/\beta)}{C + C_1 - (C_2^2/C_5) - (C_0/\beta)} \quad . \quad . \quad (45)$$

The quantity β is itself a function of α. The rigid procedure would therefore be to choose various values of α, to calculate β, and then obtain C from equation (45) and Δf from equation (44). The limits of α corresponding to the limits of C are easily found, from equation (45), to be

$$\infty > C > \frac{C_0}{\beta} + \frac{C_2}{m}\left(1 + \frac{R}{g}\frac{C_0}{L}\right)$$

$$1 < \alpha < \frac{m + 1 + (R/g)(C_0/L)}{m(C_3 + K_2/C_5) + 1 + (RC_0/gL)} \quad . \quad (46)$$

In view, however, of the fact that $\beta - 1$ is small and does not change much, it is simpler to substitute at once for α in equation (45) and to evaluate $\Delta f/f_0$.

$$\frac{\Delta f}{f_0} = \frac{1}{2}\frac{K}{K_t}\frac{\left\{C + C_2 - (C_0/\beta)\right\}(K_2 + C_5) - C_2^2}{\left\{C + C_2 - (C_0/\beta)\right\}\left\{(K_1K_2/K_t) + C_5\right\} - C_2^2} \quad (47)$$

$$\frac{\partial}{\partial C}\left(\frac{\Delta f}{f_0}\right) = -\frac{1}{2}K\left(\frac{K_2}{K_t}\right)^2 \frac{C_2^2}{\left\{C + C_2 - (C_0/\beta)\right\}\left\{(K_1K_2/K_t) + C_5\right\} - C_2^2} \quad (48)$$

This is always negative, and it increases numerically as C decreases. It is zero when C is infinite.

When C is very large, the frequency is given very nearly by

$$\frac{\Delta f}{f_0} = \frac{1}{2}\frac{K}{K_t}\frac{K_2 + C_5}{(K_1K_2/K_t) + C_5} \quad . \quad . \quad . \quad (49)$$

* See Reference (6).

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As C is decreased, the frequency increases gradually, and the rate of increase increases until the oscillations stop.

Relation between Frequency and the Various Resistances of the Circuits.—Equation (37) shows that the effect of g_2, g_3, g_4, g and R is to increase α, and therefore also to increase the frequency. This increase, moreover, becomes greater and greater as C approaches its lower limit.

Relation between Frequency and the Inter-electrode Capacities.—We have, from equation (47):—

$$\frac{\partial}{\partial C_2}\frac{\Delta f}{f_0} = -\frac{1}{2}K\left(\frac{K_2}{K_t}\right)^2 \frac{\left\{C-(C_0/\beta)\right\}^2}{\left[\left\{C+C_2-(C_0/\beta)\right\}\left\{(K_1K_2/K_t)+C_5\right\}-C_2^2\right]^2} \quad . \quad (50)$$

It is always negative, and $\left|\frac{\partial}{\partial C_2}\frac{\Delta f}{f_0}\right|$ decreases with C_2. In this respect, a large C_2 would be an advantage.

On the other hand, equation (48) shows that the greater C_2 becomes, the greater is $\left|\frac{\partial}{\partial C}\left(\frac{\Delta f}{f_0}\right)\right|$; from this point of view a large C_2 would be a disadvantage. The value of C_2 is, in general, more important in its relation to the amplitude of oscillations than in its effect on the frequency.

For the variation of frequency with respect to C_3 we find

$$\frac{\partial}{\partial C_3}\frac{\Delta f}{f_0} = -\frac{1}{2}K\left(\frac{K_2}{K_t}\right)^2 \frac{\left\{C+C_2-(C_0/\beta)\right\}^2}{\left[\left\{C+C_2-(C_0/\beta)\right\}\left\{(K_1K_2/K_t)+C_5\right\}-C_2^2\right]^2} \quad . \quad (51)$$

The frequency therefore decreases with C_3, and the larger C_3 becomes, the smaller is the modulus of the rate of decrease. Variation of K_3 has exactly the same effect.

Relation between Frequency and Air-gap.—The variation of frequency with variation of air-gap is a little more complicated than with an ordinary resonator, in which case it is given by

$$\Delta f/f_0 = \tfrac{1}{2}(K/K_t) \quad . \quad . \quad . \quad . \quad (52)$$

For the oscillator a straight line could, in theory, be obtained by plotting Δf against the right-hand term of equation (47). There would be a different graph for each value of C.

In view of the fact that, even in a resonator, the frequency changes are not governed by relation (52) above, there would be no point in carrying out the more complicated computation in the case of the oscillator.

Amplitude of Oscillations.—The amplitude is closely related to the shape of the characteristic curves of the valve. If general algebraical expressions for those characteristics were available, general expressions might be obtained for the amplitude, but the problem would present great difficulties. It is possible, however, to arrive at interesting qualitative results, by general considerations of the shape of the characteristics.

The amplitude is greatly dependent upon the capacity C. Consider a circuit like that shown in Fig. 4. The resistance of the coil is small in comparison with the anode resistance of the valve, and the mean potential at A is very nearly equal to the difference of potential V between the terminals of the battery. If the grid potential e be varied, at a frequency f, between the limits e_1 and e_2, the plate current will alternate between, say, i_1 and i_2, but will never, of course, be less than zero. If the impedance of the oscillatory circuit to current of frequency f be Z, the amplitude $v_1 - v_2$ of the alternations in potential at A will be equal to $Z(i_2 - i_1)$.

In a self-maintained oscillator, the amplitude of the oscillations, starting from a small disturbance, increases until the swing of grid potential is such that a further increase would not produce any increase in the amplitude of variation of the current.

The impedance Z of the oscillatory circuit is a maximum when C is adjusted so that $4\pi^2LCf^2$ is unity. The impedance is then equal to L/RC. Hence the amplitude is approximately inversely proportional to the resistance R of the oscillatory circuit. The law is not exact. The variations in voltage across Z are

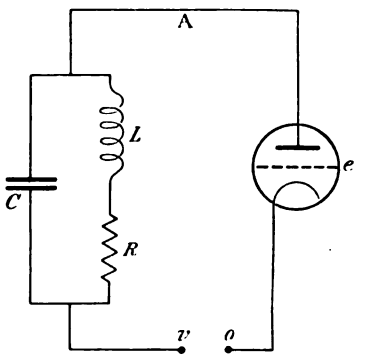

FIG. 4.

produced by variations in the plate current. The plate current is itself, however, a function of plate voltage. On the straight part of the characteristic curve of the valve,

$$v_1 - v_2 = \frac{gZ}{1+gZ}\,m(e_2 - e_1)$$

It is nevertheless always true to say that the smaller R becomes, the greater is the amplitude V. The power, however, is not, in general, a maximum at this point; it is equal to $\frac{1}{2}RC^2\omega^2V^2$, and the "useful power" is zero.

If a resistance R' be added to R, the amplitude of the oscillations will be reduced to, say, V', and the total power will be $\frac{1}{2}(R+R')C^2\omega^2V'^2$. This may, or may not, be greater than $\frac{1}{2}RC^2\omega^2V^2$. The useful power is now $\frac{1}{2}R'C^2\omega^2V'^2$. In practice the increase of resistance R' is not produced by an actual resistance added to R; it is the effective increase of resistance due to the coupling between L and the circuit in which the power supplied by the quartz oscillator is being usefully employed.

In the case of the quartz oscillator, the simple relation $LC\omega^2 = 1$, that is, $C = C_0$, cannot be used to find the value of C for maximum amplitude, because of the presence of the inter-electrode capacities.

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Suppose that, in a self-oscillatory valve circuit, the potential of the grid, before any oscillations start, is E_1 (Fig. 5). The corresponding point on the characteristic is A_1. When oscillations start, but the amplitude is still small enough, say E_2E_3, there is no change in the mean plate current. On the other hand, there is nothing to prevent the oscillations from building up, and the amplitude increases until, during part of the swing, the grid potential is positive. During this part of the swing, grid current flows and tends to lower the grid potential. During the rest of the cycle, there is no effect tending to increase the grid potential. In other words, there is a tendency for the swing of grid potential to be opposed on the positive side, but not on the negative side; as a consequence the mean grid potential falls to the left of E_1, say to E_4. The average plate current is not, in general, that corresponding to A_4, but to a point somewhere between A_4 and A_1. If, by some alteration in the valve circuits the amplitude is made to increase still more, and the maximum grid potential is, say, E_5, the effect due to grid current becomes still more pronounced, and the swing on the other side, namely, E_4E_6, is much greater than E_4E_5;

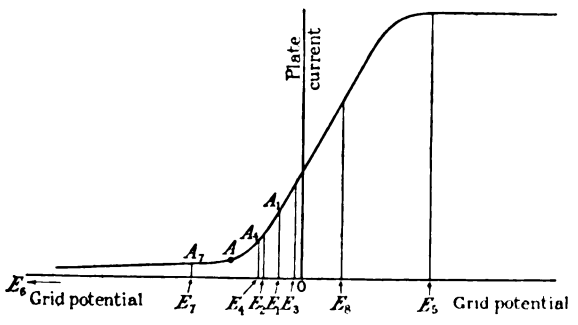

FIG. 5.

the mean grid potential decreases further to E_7, and there is a fall in the average plate current, which now corresponds to some point A between A_7 and A_4. This decrease of plate current with amplitude is well exhibited by the experimental curves in Fig. 25. The rule does not hold quite as rigidly when the resistance in the oscillatory circuit is large, because in that case the maximum amplitude is small, and the swing of grid voltage towards the positive side does not exceed, say, E_8. It may then occur that the increase in average plate current with amplitude, due to the anode rectification, more than compensates for the decrease due to the lowering of mean grid potential, itself a result of the flow of grid current.

If we ignore this slight departure from the rule, we may state that the average plate current falls as the amplitude increases. Under those conditions, the average plate conductance g will also fall with increasing amplitude, as can best be seen by the shape of the plate-volts/grid-volts characteristics of a valve, for various plate currents.

Consider now expression (41), and suppose that the capacity is at first very large; no oscillations can take place because the expression is positive. The plate current is fixed solely by the plate, filament, and grid voltages. We may suppose, for the sake of argument, that there is a grid leak; the potential of the grid is then equal to the potential of the filament; denote the corresponding plate current and plate conductance by I_0 and g_0. Let C be decreased gradually. When it reaches the limit given by equation (42), expression (41) is just zero, and oscillations are on the point of starting. The value of g to be used in equation (42) is therefore g_0. Suppose C to be decreased further. If the value of g were still g_0, expression (41) would be negative. Oscillations therefore build up, and the amplitude increases, until g reaches the value such that expression (41) becomes equal to zero. This new value of g will depend upon the setting of C, and the amplitude will be some function of $1/g$. It might be possible to find the amplitude in terms of the average g from a complete set of characteristics, but the process would, undoubtedly, be complicated. It will, nevertheless, be interesting for us to plot the value of $1/g$ against C, and to compare the curve thus obtained with the amplitude curves obtained experimentally. The curves must not, of course, be expected to coincide. All that can be expected is that (d/dC) should have the same sign for both curves.

When the value of C is that given by equation (42), oscillations are just on the point of starting, the amplitude is zero, and the plate conductance is g_0. It is therefore convenient to plot $(1/g) - (1/g_0)$, or $r - r_0$, against C, where r, the reciprocal of g, is the mean internal plate resistance of the valve.

The effect of some of the circuit variables on the amplitude can be predicted without actually plotting the curves.

The terms in S, g_1, g_2, g_3, g_4 and R, all tend to raise the value of g which makes expression (41) equal to zero. All the damping terms are, therefore, in general, detrimental. It may occur, however, that by a judicious choice of the grid leak g_4 the mean grid potential becomes such that oscillations take place about a more favourable point on the characteristic curve of the valve. Under these conditions the effect of g_4 would be beneficial.

The conductance g_3 between anode and filament may be considered as being made up of two parts: the first, g_3', a small direct-current conductance due to want of perfect insulation, and the second, g_3'', due to the dielectric losses in the glass and holder. This latter component, by far the larger of the two, merely increases the damping and slightly reduces the amplitude of oscillations; the component g_3' tends to equalize plate and grid potential, and therefore to stop the oscillations. It is, however, very small, being probably less than 10^{-8} mho in most valves.

The third term of expression (41) contains as a factor $C_2/\{K_1 + C_5(K_1/K_2)\}^2$.

When C_2 is zero, expression (41) will be positive, and no oscillations can take place. In the same way, large values of C_2 make the third term small numerically, and the expression will again be positive. There will therefore be, for each value of C, an optimum value of C_2.

The inter-electrode plate-to-grid capacity C_3, and the stray capacity K_3 in parallel with it, increase C_5, and therefore increase the term in S in expression (41). Their presence is, therefore, always detrimental; hence

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the use, by J. R. Harrison, of a screened plate, to do away with the capacity C_3. Hence also the importance of keeping as low as possible the stray capacity K_3 between the electrodes and leads of the resonator.

The term in S is multiplied by

$$\{K_1 + C_5(K_l/K_2)\}^2$$

The amplitude obtainable with a small air-gap will thus be larger than the amplitude with a large air-gap, because the smaller the air-gap the smaller is (K_l/K_2).

The value of g obtained by equating expression (41) to zero is independent of the initial value g_0, other things being equal. In a low-impedance valve, however, the swing of plate current required to produce a given change in the mean plate conductance is larger than in a high-impedance valve. Hence, the lower the normal plate impedance r_0, the greater will the amplitude be.

The third term in expression (41) contains the amplification constant m as a factor. The greater m, the smaller will be the value of g which makes the expression equal to zero; a large amplification factor is therefore beneficial, other things being the same.

The analysis of expression (41) is greatly simplified by supposing that

$$C_2 + C - \frac{C_0}{\beta} - \frac{C_2^2}{C_5}$$

and

$$C_2 + C - \frac{C_0}{\beta} - \frac{C_2^2}{(K_1K_2/K_l) + C_5}$$

are nearly equal. In numerical computation this supposition is not even necessary.

Writing, for brevity, $C - (C_0/\beta) = x$, we can put expression (41) into the form

$$\Sigma\left[(a_1x+a_2)^2+(a_3x+a_4g+a_5)^2\right]+a_6+a_7g-a_8gx \quad . \quad (53)$$

in which the coefficients a_1, . . . etc., are all positive. The Σ includes four similar terms, which contain, as factors, S, g_1, g_3, and $g_2 + g_4$ respectively; some of the coefficients a_1, a_2, . . . etc., may be zero. We see at once that there cannot be any maintained oscillations unless x be positive. The expression can then be put into the simpler form

$$\phi = x^2 + Ax + B + Cg^2 + Dg - Exg \quad . \quad (54)$$

where the coefficients A, B, C, D and E are all positive.

Since x is positive, oscillations cannot be maintained unless

$$Ex > D$$

This, however, is not the only condition, because some positive values of x might give imaginary roots for g in the equation $\phi = 0$. We therefore require the further condition

$$\phi_2 = (Ex - D)^2 - 4C(x^2 + Ax + B) \lesseqgtr 0 \quad . \quad (55)$$

Since Ex is positive and greater than D, this condition requires that

$$E^2 > 4C$$

Whenever the inequality (55) is satisfied, the two values of g given by the equation $\phi = 0$ are real. They are, moreover, both positive, since their sum and their product are positive.

The differential coefficient of ϕ_2 is

$$2E(Ex - D) - 4C(2x + A)$$

But we also have

$$(Ex - D)^2 - 4C(x^2 + Ax + B) \lesseqgtr 0$$

$$\therefore \quad Ex(Ex - D) - 4C(x^2 + Ax + B) > 0$$

$$2Ex(Ex - D) - 4C(2x^2 + 2Ax + 2B) > 0$$

$$2E(Ex - D) - 4C(2x + A) > 0$$

The differential coefficient of expression (55) is thus positive for all values of x over which oscillations may take place. Expression (55) therefore increases with x.

Denote by ξ_2 the larger of the two roots of the equation $\phi_2 = 0$. For the maximum amplitude, that is, for the minimum value of g with respect to x or C, we have the equations

$$\phi = 0 \quad \text{and} \quad dg/dx = 0$$

But

$$\frac{\partial\phi}{\partial x} + \frac{\partial\phi}{\partial g}\frac{dg}{dx} = 0$$

The condition for a minimum g is therefore

$$\left.\begin{aligned}\phi &= 0\\ \partial\phi/\partial x &= 0\end{aligned}\right\} \quad . \quad . \quad . \quad . \quad . \quad (56)$$

The second of these equations is

$$2x + A - Eg = 0 \quad . \quad . \quad . \quad (57)$$

Substituting for g in the first, we find

$$(Ex - D)^2 - 4C(x^2 + Ax + B) = BE^2 + A^2C + ADE + D^2 - 4CB \quad (58)$$

But since $E^2 > 4C$, the right-hand term of expression (58) is always positive, and the value ξ_1 of x given by expression (58) is greater than ξ_2. The amplitude therefore reaches a maximum before the point of cut-off.

In general, the two equal values of g corresponding to ξ_2, namely $\frac{E\xi_2 - D}{2C}$, are smaller than g_0, and the corresponding value of r is larger than r_0. As soon as C is reduced to $\xi_2 + (C_0/\beta)$, oscillations stop suddenly. For this reason the value of C corresponding to ξ_2 will be called the "cut-off point." If the damping terms were large, ξ_2 would be large, and the corresponding value of r might be less than r_0. In such a case the curve of r against C would cut the line $r = r_0$ at some point ξ_3 between ξ_1 and ξ_2. If C were decreased gradually, the amplitude would rise to a maximum and then decrease gradually, and become zero at $C = (\xi_3 + C_0/\beta)$, but there would be no sudden cut-off.

Two other interesting points are obtained by substituting g_0 for g in equation (54) and solving for x.

Unless the damping terms be very large, Eg_0 is greater than A and there are two positive roots. If A were greater than Eg_0, no oscillations could ever start of themselves. It is easily seen by inspection that the smaller the damping terms, the smaller will be the product of the roots and the larger will be the difference

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decrease the frequency. This decrease is, moreover, greater and greater as C approaches its upper limit.

Relation between Frequency and the Inter-electrode Capacities.—We have, from equation (70),

$$\frac{\partial}{\partial C_3}\frac{\Delta f}{f_0} = -\frac{1}{2}K\left(\frac{K_2}{K_t}\right)^2 \frac{\{(C_0/\beta) - C\}^2}{\left[\{(C_0/\beta) - C - C_3\}\{(K_1K_2/K_t) + C_5\} + C_2^2\right]^2} \quad . \quad . \quad . \quad . \quad . \quad (75)$$

This is always negative, and therefore the frequency decreases with increase of C_3.

Similarly, we have

$$\frac{\partial}{\partial C_2}\frac{\Delta f}{f_0} = -\frac{1}{2}K\left(\frac{K_2}{K_t}\right)^2 \frac{\{(C_0/\beta) - C - C_3\}^2}{\left[\{(C_0/\beta) - C - C_3\}\{(K_1K_2/K_t) + C_5\} + C_3^2\right]^2} \quad . \quad . \quad . \quad . \quad . \quad (76)$$

which shows that the frequency decreases with increase of C_2, but that the rate of decrease decreases.

Amplitude of Oscillations.—Expression (71) shows that, when C_3 is zero or infinite, no oscillations can take place. There is, therefore, for each value of C, a particular optimum value of C_3 which makes g a minimum when expression (71) is equated to zero and solved for g.

The capacity $C_2 + K_3$, in parallel with the quartz resonator, is always detrimental.

Curves II, IV and VI in Fig. 6 show the excess mean internal resistance, $r - r_0$, for various values of C.

When C is very small—it can never be zero, because it includes the inter-electrode capacity C_1 and the self-capacity of the coil L—there may or may not be oscillations, depending upon whether C is above or below the limiting value given by equation (73). When C is increased past this value, oscillations start at a very small amplitude, which increases slowly at first, then more rapidly, until it reaches a maximum. Past this point the amplitude decreases very rapidly, until a certain value of C is reached at which the oscillations stop suddenly. The capacity at cut-off is very near to the capacity for a maximum, especially if the damping terms, that is, the terms containing S, g_1, g_2, g_3, g_4 and R, are small.

(2) EXPERIMENTAL INVESTIGATION.

Description and Behaviour of the Quartz Disc.

Only one piece of quartz was used throughout the experiments. It was a disc 4·85 cm in diameter and 0·752 cm thick. The electric axis was perpendicular to the plane of the disc, and the only mode of vibration studied was that which took place along the electric axis.

According to Hund's empirical formula, the frequency should have been equal to 287/0·752, or 382 kilocycles per sec.

The piece of quartz, however, gave two strong responses at frequencies of approximately 389 and 385 kilocycles per sec. respectively. Besides these, there were several weak responses at frequencies not very different from the above.

The response at 389 kilocycles per sec. was a little stronger than that at 385. When the quartz electrodes were connected to the grid and plate of the valve, oscillations were, in general, maintained only at 389 kilocycles per sec.; when the filament-to-grid connection was used, and the capacity in the oscillatory circuit was very small, no oscillations were maintained. As the capacity was gradually increased, oscillations started, and their amplitude gradually increased to a maximum; the frequency of these oscillations was approximately 389 kilocycles per sec. A further slight increase in the capacity would alter the circuit conditions, and the oscillations stopped. At this point the quartz started oscillating in the other mode and, as the capacity was further increased, the amplitude increased, and then decreased almost abruptly to zero at the value of capacity for which the circuit conditions for the maintenance of oscillations at 385 kilocycles per sec. were no longer fulfilled. The difference between the two limiting capacities was, of course, approximately equal to (2 × 4)/390, or 2 per cent of the total capacity.

Under certain conditions it was possible to obtain the two oscillations simultaneously: an audible note of approximately 4 000 cycles per sec. was then produced.

Air-Gap and Electrodes.

For the air-gap and electrodes use was made of an apparatus designed by Dr. Dye for the interferometric examination of vibrating quartz plates. The apparatus as used for the present purpose is shown in Fig. 7. The

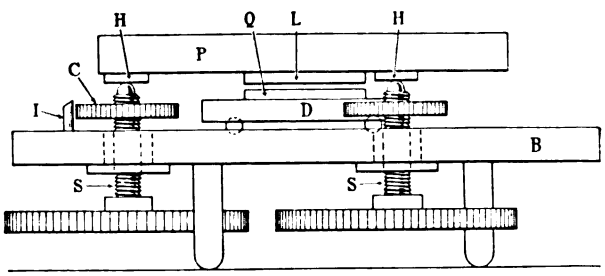

Fig. 7.

quartz disc Q rests by its own weight upon a brass electrode D, which, in turn, is supported upon three small steel balls (by hole, slot and plane location) upon the main base-plate B. The base-plate carries 3 well-fitted levelling screws S with ball ends and knurled ebonite heads. Each screw is fitted with a graduated circle C. The pitch of the screw is ½ mm, and the circumference of the circle is divided into 50 equal parts, of which a tenth part can be estimated without much difficulty. This is equivalent to 0·0001 cm. The position of the circle is read by means of the index I, fixed to the base-plate. The screws support a glass plate P, of which the lower surface is plane polished. To this lower surface are wrung 4 polished brass discs, 3 of which, H, rest directly on the screws; the fourth, L, constitutes the upper electrode; its diameter exceeds that of the quartz disc by 2 mm. To this disc is soldered

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a thin copper wire which connects it to a terminal of the apparatus; the second terminal is in electrical connection with the base-plate. When the quartz is used as a resonator, the insulated terminal of the main condenser is connected to the disc, because the capacity to earth of the disc is much smaller than the capacity current of audible frequency. The 3 screws are lowered until the ball ends are out of contact with the discs H. The upper electrode L is then resting on the upper face of the quartz, and the air-gap is zero. Each of the screws is adjusted until the slightest further upward motion would bring them into contact with the discs H,

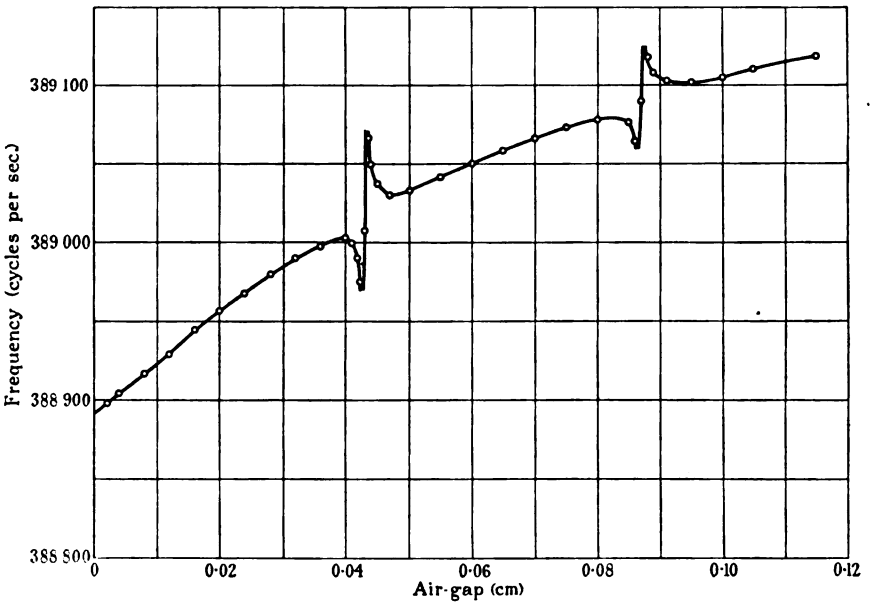

FIG. 8.—Response frequency of the resonator at various air-gaps.

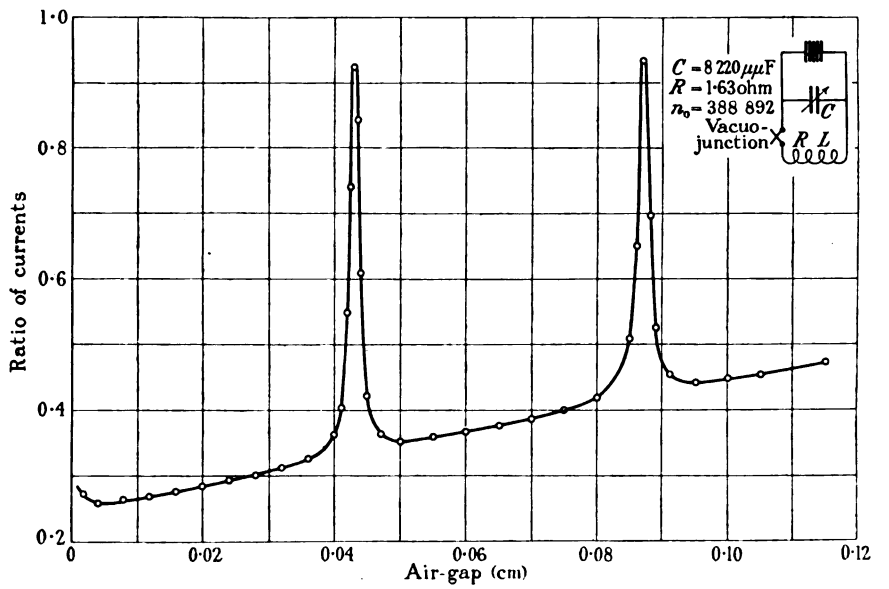

FIG. 9 —Ratio of minimum current to current when the quartz is not vibrating.

to earth of the base-plate. For the same reason the disc is always connected to the grid when the quartz is used with a valve.

In order to obtain the screw readings for zero air-gap, the discs H are connected together and to one terminal of a telephone. The base-plate is connected to the other terminal of the telephone through a source of alternating and the position of the circles is noted. Contact is, of course, indicated by the telephone.

ELECTRICAL CONSTANTS OF THE QUARTZ.

The constants were measured by Dye's method. The quartz electrodes are connected to the terminals of the condenser of an oscillatory circuit, and the inductance

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coil of the circuit is loosely coupled to a source of alternating power at approximately the quartz frequency. The frequency is varied and the current in the oscillatory

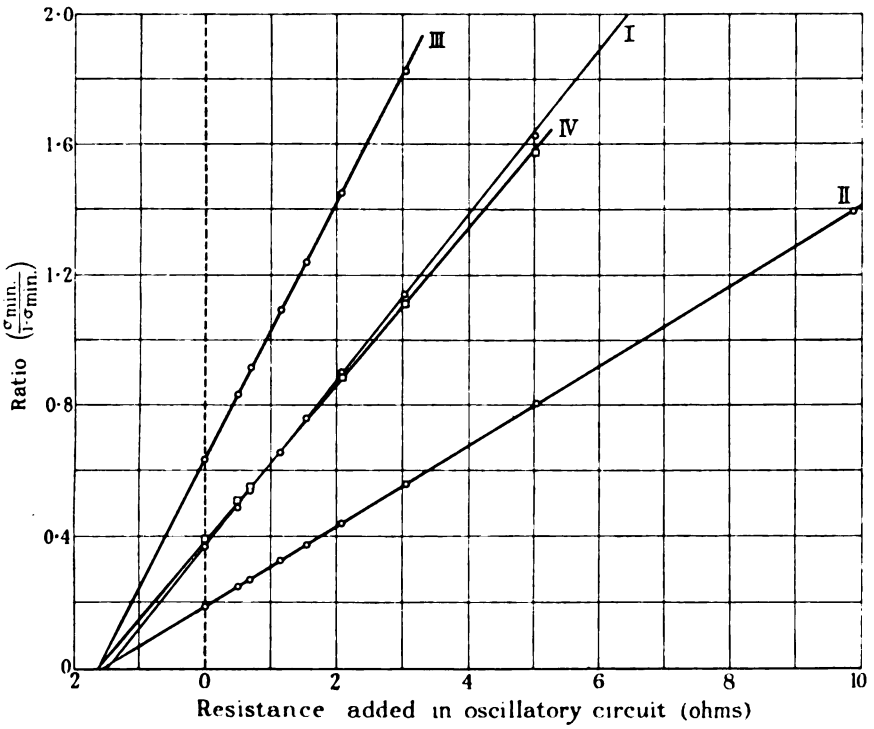

FIG. 10.—Determination of resistances of the circuits of the resonator.

circuit is measured by means of a thermo-junction placed in series with the inductance coil. The reader

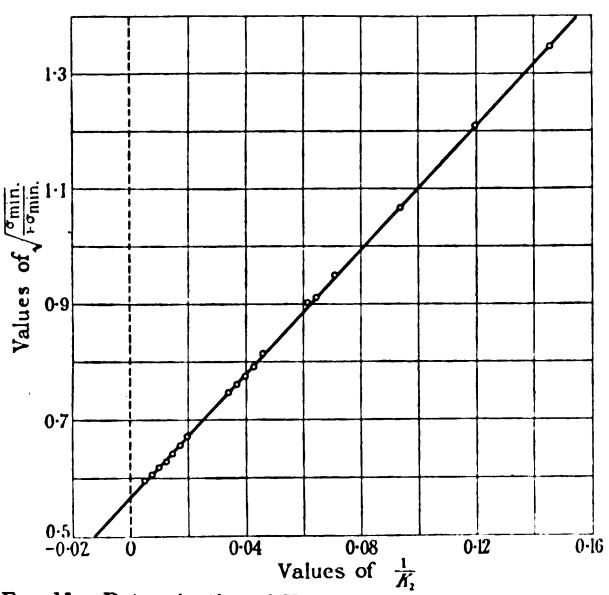

FIG. 11.—Determination of K_1, and of S, the damping of the resonator.

who desires more information on this method should consult Dye's paper.*

* See Reference (4).

This method of analysis necessitates a variation of air-gap and a knowledge of the corresponding capacity K_2. This capacity is calculated by the formula $S/3{\cdot}6\pi g$ micro-microfarads, where S is the area of the surface of the quartz, and g the length of the air-gap in cm. With the dimensions given above, $K_2 = 1{\cdot}633/g$ micro-microfarads.

Fig. 8 shows the frequency for various air-gaps, and Fig. 9 the corresponding value of the minimum current in the inductance coil. The large increase of damping at air-gaps of 0·043 cm and 0·087 cm is due to the resonance of the air in the gap. The corresponding velocity of propagation of the waves is 34 200 cm per sec. The average room temperature was 16·5° C.

Fig. 10 gives the resistances of the two circuits used. The capacities of the two circuits were obtained by direct reading of the condensers. Fig. 11 gives $K_1 = 9{\cdot}6$ $\mu\mu$F, and also the damping S of the quartz.

This damping S depends upon the mode of support of the quartz and upon the atmospheric conditions. It is therefore necessary to measure it anew whenever experiments are being carried out on the quartz. Once K_1 has been obtained, S can be calculated from a measurement of the minimum current for a single value of the air-gap K_2. The value of S varied between 350 and 500 ohms.

Resonance curves like Fig. 12 were obtained for four different air-gaps. They give the value of $f_2 = SK\omega$, and, since S can be obtained from the minimum σ as explained above, the value of K is found. This is equal to 0·018 $\mu\mu$F.

The graph of frequency against $1/(K_1 + K_2)$ should, according to the formulæ, be a straight line, and would

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give a means of checking the value of K. In practice, however, this graph is not a straight line, as Fig. 13 shows. This want of agreement occurs in all resonators,

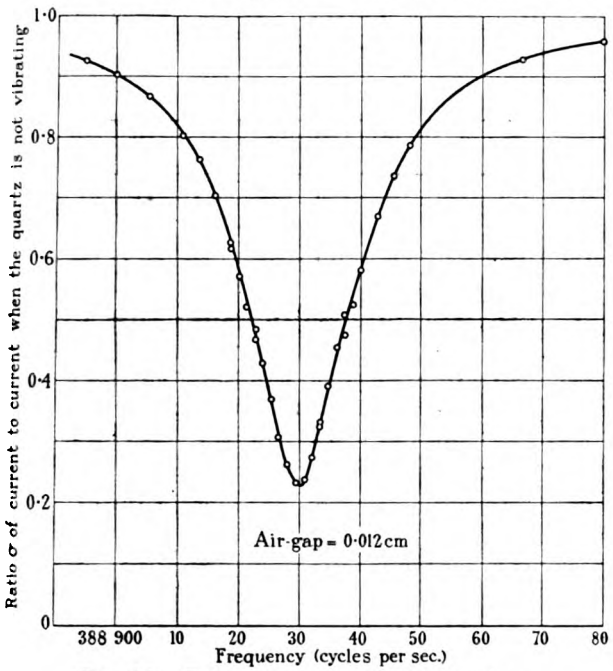

FIG. 12.—Resonance curve of the resonator.

and deserves careful consideration. We cannot, however, dwell upon this point here, because we are mainly concerned with quartz oscillators.

The ratio of K_1 to K is 530, instead of the theoretical value 140. The value of K_1, calculated from a permittivity of 4·55, is 9·9, which does not differ much from the experimental figure 9·6 found above.

The investigation on the quartz used as an oscillator, in connection with a valve, was carried out for only one value of the air-gap, namely 0·012 cm. The corresponding value of K_2 is 136 $\mu\mu$F, and the value of (K_2/K_t) is 0·935.

The stray capacity K_3 between the two electrodes and their leads was deduced from measurements of the total capacity between the ends of those leads for various frequencies on each side of the resonance frequency. The value was 15 $\mu\mu$F.

The curve in Fig. 12 gives the response frequency of the resonator at an air-gap of 0·012 cm and a room temperature of 20·5° C. It is 388 930 cycles per sec. All the frequencies given in this paper have been reduced to the same room temperature, namely 20·5° C.

Fig. 8 also gives f_0, the response frequency for zero air-gap; it is 388 892 cycles per sec.

If the frequency at an air-gap of 0·012 cm were calculated from f_0 by the formula

$$f - f_0 = (f_0/2)\,(K/K_t)$$

the result would be 388 916 cycles per sec. instead of the experimental value given above, namely 388 930 cycles per sec.

An analysis similar to the above was also made for the response occurring at a frequency of approximately 385 kilocycles per sec., with a view to explaining why the quartz disc oscillated more readily at 389 than at 385 kilocycles per sec. The results of the two measurements are given in Table 2. The two values of S were obtained at a few minutes' interval, and are therefore

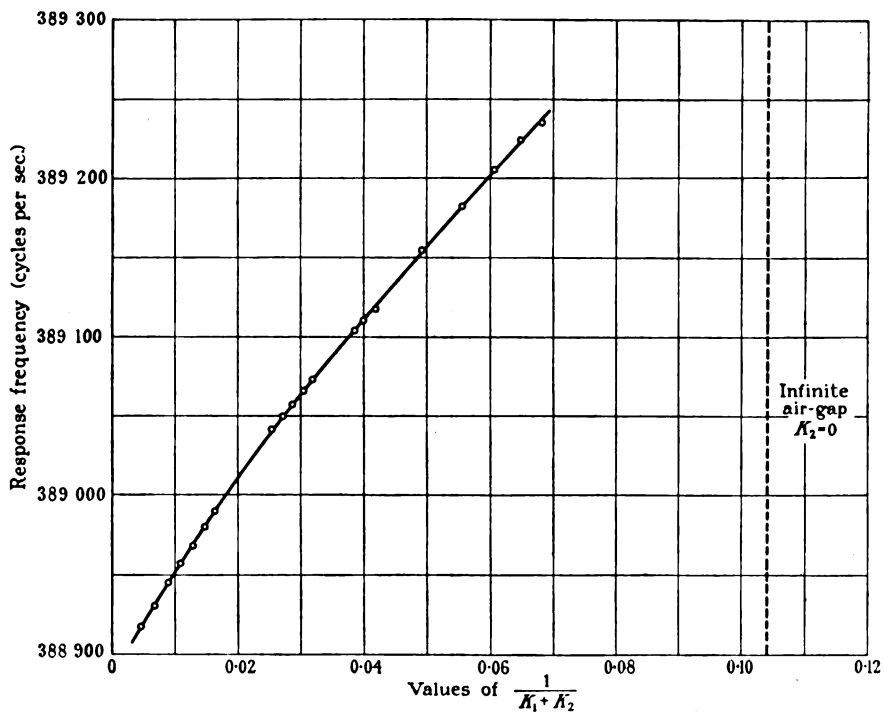

FIG. 13.—Relation between response frequency of the resonator and air-gap.

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comparable. They would not be comparable if they had been obtained under different atmospheric conditions.

The reason why the quartz disc oscillates more readily at the higher frequency is now obvious from Table 2.

TABLE 2.

	First response	Second response
K_1 ($\mu\mu$F)	9·6	9·4
K ($\mu\mu$F)	0·018	$0{\cdot}013_3$
S (ohms)	400	540
f_0 (cycles per sec.) ..	388 890	385 210
Logarithmic decrement ..	55×10^{-6}	55×10^{-6}

The logarithmic decrement is approximately the same in both cases, but its constituent parts are different. In one case S is small and K is large; this is equivalent to saying that the damping is small and that a relatively large proportion of the quartz takes part in the vibration. For the second response, S is rather larger, and K is smaller. The damping is therefore larger, and the vibrations take place over a smaller portion of the face of the disc. Alternatively, a small K might indicate that the vibrations in certain parts of the disc are opposite in phase to the vibrations in other parts.

VALVES.

For most of the tests a D.E.R. valve was used. The values of g and m, which occur in the equation $i = g(v + me)$, were determined from the characteristics. They are given in Table 3.

TABLE 3.

D.E.R. Valve No. 3343.

Plate current	Plate conductance g
mA	mho × 10—6
0·5	27
1·0	31
2·0	40
3·0	41·7

Amplification factor $m = 8{\cdot}6$

The tests on the quartz oscillator were made with 100 volts between the plate and the filament, and 2 volts on the filament. Under those conditions the plate current was 3 mA when the grid was kept at filament potential by means of a grid leak or in any other way.

In some cases a D.E.P. 610 valve was used for comparison with the D.E.R. valve. The filament voltage was 6·4 and the plate voltage 100. The amplification factor m was approximately 8, and the conductance g was as given in Table 4. With the grid at filament potential, the plate current was 15mA for a plate voltage of 100.

Inter-electrode Conductances and Capacities of the D.E.R. Valve.—The inter-electrode capacities and conductances of the valve are represented diagrammatically by Fig. 14. The three pairs of quantities were measured

TABLE 4.

Valve D.E.P. 610.

Plate current	Plate conductance g
mA	mho × 10—6
5	290
10	260
15	200

on the Schering radio-frequency bridge by a method analogous to that described two years ago by Hartshorn.* There were slight modifications because the radio-frequency bridge and the bridge used by Hartshorn are not identical. The results obtained are given in

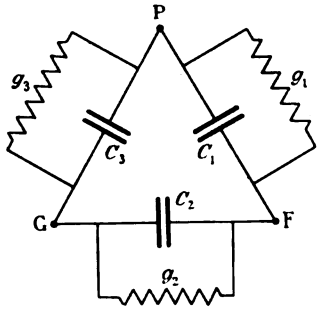

FIG. 14.

Table 5. The measurements were taken with the filament hot, and do not differ much from the results obtained with the filament cold.

TABLE 5.

D.E.R. Valve.

$\mu\mu$F	mho × 10—6
$C_1 = 4{\cdot}0$	$g_1 = 0{\cdot}2$
$C_2 = 6{\cdot}4$	$g_2 = 0{\cdot}5$
$C_3 = 2{\cdot}4$	$g_3 = 0{\cdot}05$

The relatively large capacity between grid and filament is due to the extra terminals between which the grid leak is inserted. The capacities are accurate to within 0·2 $\mu\mu$F, but the accuracy of the conductance figures is much less; it is probably 30 per cent, and might be as low as 50 per cent. To obtain accurate values, the valve should have been screened.

GRID RESISTANCES.

The grid resistances used were Loewe high-vacuum resistances, type FZ 128, which are platinized glass rods enclosed *in vacuo*.

The resistances were measured on a Wheatstone bridge and also on the Schering radio-frequency bridge, at a frequency of approximately 400 000 cycles per sec.

* See Reference (7).

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The results are given in Table 6.

For the radio-frequency measurements it was assumed that the grid resistances were equivalent to a pure resistance in parallel with a pure capacity. In the radio-frequency measurements the amplitude of the alternating voltage applied to the resistances was 4 volts; measurements were also taken with an amplitude of

TABLE 6.

Loewe Grid Resistances.

Nominal resistance	Conductance		Parallel capacity at 400 000 cycles per sec.
	To direct current	At 400 000 cycles per sec.	
megohms	mho × 10⁻⁶	mho × 10⁻⁶	μμF
3	0·34	0·36	0·1
2	0·43	0·45	0·2
1	0·87	0·8	0·1
0·5	1·78	1·7	0·1
0·25	4·4	4·5	0·2
0·1	11·1	11·2	0·2
0·075	11·9	12	0·2
0·05	17·0	17	0·3

10 volts on the 1-megohm grid leak, and no difference could be detected.

The accuracy of the radio-frequency measurements is well within 5 per cent.

OSCILLATORY CIRCUITS.

The circuit used during the greater part of the investigation had an inductance of 160 μH, a self-capacity of 20 μμF and a total resistance of 2·08 ohms. In series with the condenser was a vacuo-junction of resistance 1·12 ohms, so that the resistance of the coil, leads and condenser was approximately 0·96 ohm.

In some of the experiments another circuit was used. Its inductance was 325 μH and its self-capacity 25 μμF. The total resistance was 4·3 ohms, 1·1 ohms being due to the vacuo-junction and 3·2 ohms to the coil, leads and condenser. The inductances were measured in each case by the capacity-variation method at frequencies in the neighbourhood of 400 000 cycles per sec., and the accuracy is probably better than 0·5 per cent.

With the D.E.P. 610 valve the currents in the condenser were large, and a thermo-junction with constantan heater was used instead of the vacuo-junction. The heater resistance was 0·61 ohm both on direct current and at a frequency of 390 000 cycles per sec. The total resistances of the two circuits were then 1·57 ohms and 3·8 ohms respectively.

Throughout the paper the term "total capacity in the oscillatory circuit" includes the self-capacity of the circuit and the parallel inter-electrode capacity C_1 between the plate and the filament.

Complete Circuits.—The circuits used are shown in Figs. 15 and 16. The telephones had a total resistance of 300 ohms to direct current, and a by-pass condenser of 0·02 μF was connected between the filament and the screen of the condenser of the oscillatory circuit. Resistance could be added at B in the oscillatory circuit.

Measurement of Alternating Voltage across the Condenser.—The heater of the thermo-junction was in series with the condenser, and not in series with the coil, because the plate current also flows through the coil and would vitiate the readings. The R.M.S. voltage across the coil is then given by the well-known formula $V = I/(C\omega)$; the current I is itself obtained from the reading of the galvanometer. Each thermo-junction

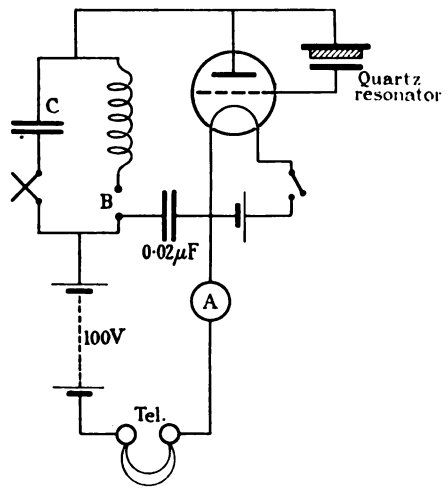

FIG. 15.—Plate-to-grid connection.

was calibrated at the outset of the investigation, and the heater current was found to be proportional to the galvanometer deflection, as read on a square-law scale. The galvanometer is a moving-coil instrument and the scale is at a distance of 150 cm from the mirror.

The instrument repeats with an accuracy of about 0·2 per cent, but the absolute value of the voltage across the condenser is only known to the accuracy of

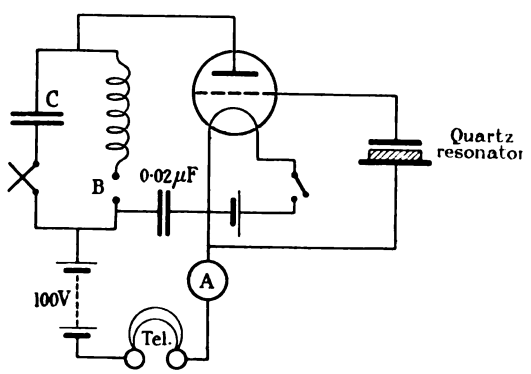

FIG. 16.—Filament-to-grid connection.

the standard milliammeter used in the calibration of the thermo-junctions. This was about 0·5 per cent. It is clear that a high degree of accuracy in the absolute value is not necessary in the present case.

Measurement of Frequency.—The inductance coil of the quartz oscillator was very loosely coupled to the grid coil of a sensitive tuned-anode amplifier. The output coil of an auxiliary valve-oscillator was also made to induce in the amplifier coil, but was given by

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trial a position such that its coupling to the inductance coil of the quartz oscillator was as small as possible. The auxiliary generator was provided with variable condensers which ensured smooth and continuous variation of frequency. One of the condensers gave a variation of 2·65 cycles per sec. for 1 $\mu\mu$F, and it was a simple matter to read to within 0·1 $\mu\mu$F.

The oscillations from the quartz oscillator and the oscillations of approximately the same frequency from the auxiliary generator are received and amplified by reading of the condenser of the auxiliary generator must be altered to restore coincidence between the beat-tone frequency and the tuning-fork frequency. Variations in the frequency of the quartz oscillator can thus be estimated from the readings of the condenser of the auxiliary generator.

This condenser is calibrated by taking readings corresponding to beat tones of various frequencies, the frequencies being those of known tuning-forks.

The absolute frequency is given by the multi-vibrator

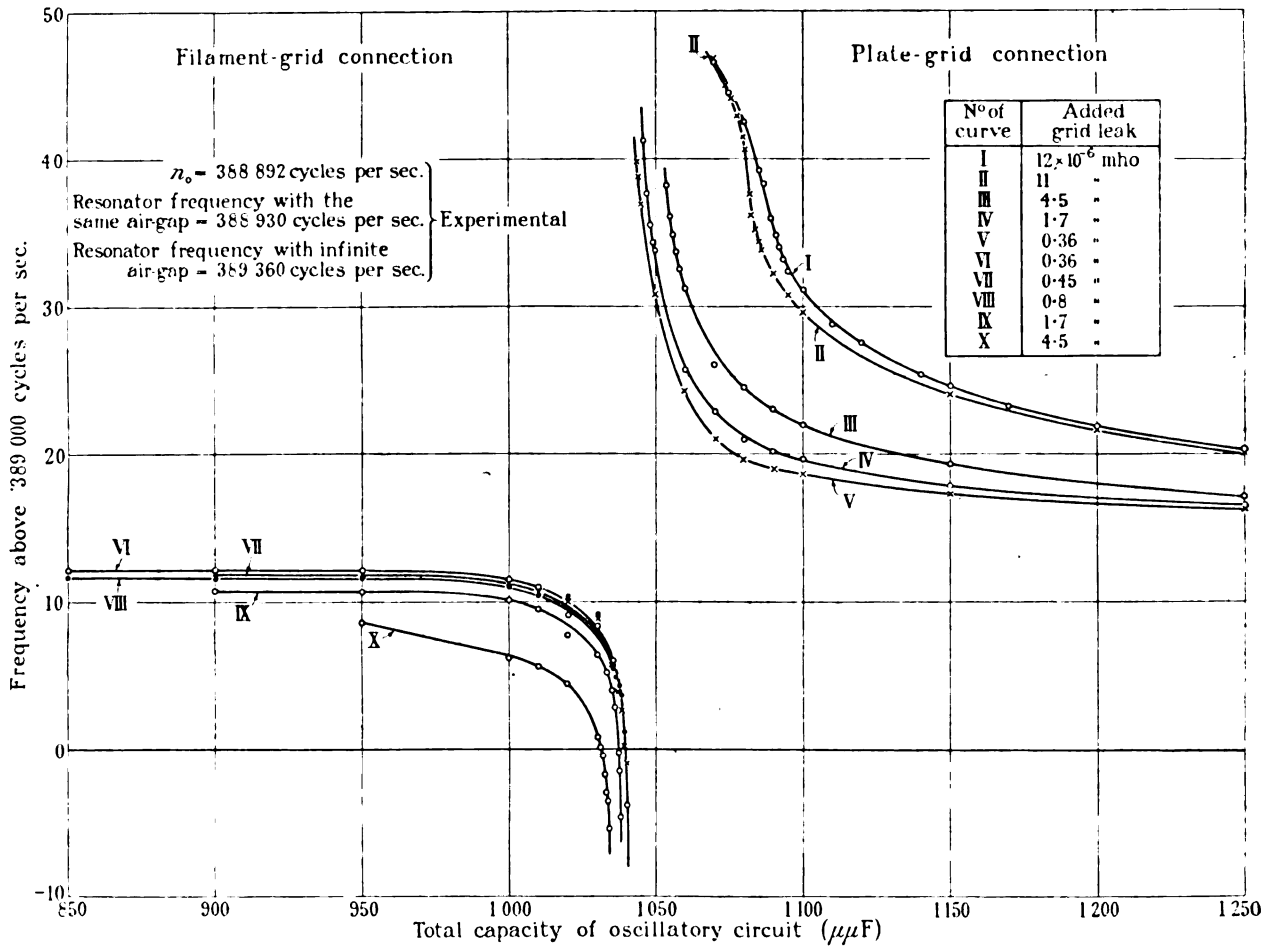

FIG. 17.—Relation between the frequency and capacity of the oscillatory circuit for various grid leaks.
Valve: D.E.R. H.T. voltage = 100; L.T. voltage = 2; R = 2·1 ohms; L = 160 μH; air-gap = 0·012 cm.

the first two stages of the tuned-anode amplifier; if the two radio frequencies are nearly equal, an audible beat tone is produced which is amplified in the last two stages of the amplifier and reproduced by the loud-speaker.

In order to compare the two radio frequencies, the condenser of the auxiliary generator is varied until the frequency of the beat tone coincides with the frequency of a tuning-fork. In the present case a tuning-fork of 1 000 cycles per sec. was used.

When the frequency of the quartz oscillator is varied by an alteration in one or more of the constants of its circuit the frequency of the beat tone alters, and the wavemeter. This instrument induces known frequencies, multiples of 20 000, in the tuned-anode amplifier. Any one of those frequencies can be reinforced at will by tuning the grid circuit of the amplifier. The condenser of the auxiliary generator is adjusted until its frequency coincides with the multi-vibrator frequency. Equality is indicated by a characteristic beat. Other characteristic beats indicate equality between the auxiliary generator frequency and multiples of 1 000, and also multiples of still smaller numbers. In the present case it was quite easy to obtain calibration points 500 cycles per sec. apart.

The variations in frequency due to variations in the

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temperature of the quartz are much larger than many of the variations that we are concerned with here. It was found, moreover, that the temperature of the quartz was a function of the amplitude of its vibrations. Such effects were allowed for by returning to a standard state between successive readings. In parallel with the calibrated condenser A of the auxiliary generator was a small condenser B by which any frequency drift could be corrected. The following example will illustrate the mode of procedure: Suppose it is desired to obtain a relation between the frequency of the quartz oscillator and the reading C of the condenser in the oscillatory circuit. Condenser C is set at a standard value C and condenser A at a standard value A. The other condensers of the auxiliary generator are varied until the beat tone is, say, 1 000 cycles per sec. Condenser C is moved to a value C_1 and A is altered until the beat tone is again 1 000, to A_1, say, giving the frequency difference looked for. Condensers A and C are then brought back to A and C, and the beat tone should then be 1 000. If there has been a small drift, it is taken on B, and the whole process is repeated for other values C_2, C_3, etc., of the condenser C. Variations of the quartz circuit are measured in the same way, and the complete set of results afterwards correlated. Only a single measurement of the absolute frequency is therefore necessary. In practice several were taken to serve as checks.

Relation between Frequency and Total Capacity in the Oscillatory Circuit.—A set of curves, reproduced in Fig. 17, was obtained by the method described in the previous Section. Simultaneous readings of the condenser of the auxiliary generator, of the deflection of the galvanometer spot and of the milliammeter in the plate circuit, were taken, and, for correcting drifts, readings were frequently taken for one and the same setting of the condenser in the quartz circuit. Curves were obtained with grid leaks of widely different values; the actual figures are given in the table in Fig. 17.

The capacity in Fig. 17 is the total capacity C, obtained by adding to the condenser reading 20 $\mu\mu$F for the self-capacity of the oscillatory circuit, 4 $\mu\mu$F for the inter-electrode plate-to-filament capacity C_1 and 6 $\mu\mu$F for the stray capacity between the plate lead and the leads of the by-pass condenser.

The shape of the curves confirm in a general way the conclusions arrived at by theoretical considerations on pages 46 and 51; namely, that

(i) With the resonator connected between the plate and the grid:

- (*a*) The frequency increases when C decreases;
- (*b*) The rate of variation increases as C decreases;
- (*c*) The greater the conductance of the grid leak, the higher the frequency; and
- (*d*) The rate of variation of frequency for a given value of C is greater with large grid conductances than it is with small grid conductances.

(ii) When the resonator is connected between the grid and the filament:

- (*a*) The frequency decreases when C increases;
- (*b*) The rate of variation of frequency increases as C increases;
- (*c*) The greater the conductance of the grid leak, the lower the frequency; and
- (*d*) The rate of variation of frequency for a given value of C is greater with large grid conductances than it is with small grid conductances.

The curves obtained when there was no added grid resistance do not fit in well with the others. The reason is that when no grid leak was used, simultaneous oscillations were produced at two different frequencies, as described on page 52, even with the plate-to-grid type of connection. It is quite conceivable that the mode of vibration under investigation, giving the frequency of approximately 389 kilocycles per sec., would be affected by the other, of frequency approximately 385 kilocycles per sec.

When the resonator is connected between the grid and the plate, the frequency without a grid leak should be slightly lower than the frequency with a grid leak. In the present case, however, it was found to be 0·9 cycle per sec. higher than with a grid leak of 3 megohms.

Another reason why it is better to consider only those curves obtained with grid leaks is that the added grid conductance, even though small, fixes the grid potential. With the D.E.R. valve used, the plate current was only 2·7 mA when there was no grid leak and when no oscillations were being maintained; with a grid leak, the corresponding value was 3·0 mA.

The next step is to calculate the frequency at various condenser readings and to compare the results with the curves of Fig. 17. This has been done for both types of connection for the case when the added grid conductance was $0{\cdot}36 \times 10^{-6}$ mho.

Consider first the plate-to-grid type of connection. The general procedure is to select particular values of α between the limits $\alpha = 1$ and $\alpha = 1{\cdot}301$, and to calculate β from equation (35) and then C from equation (45). The frequency is given by $f = f_0(\frac{1}{2} + \frac{1}{2}\beta)$.

The computation is, however, greatly simplified by making use of the fact that, as α varies between its limits 1 and 1·3, as given by equation (46), β only varies between the values $1 + (6 \times 10^{-4})$ and $1 + (7 \times 10^{-4})$. Regarding first β as constant and equal to the latter value, the frequency is calculated from equation (44) or (47).

The circuit variables were as follows:—

$$C_2 = 6{\cdot}4\ \mu\mu\text{F}$$
$$C_3 = 2{\cdot}4\ \mu\mu\text{F}$$
$$K_3 = 15\ \mu\mu\text{F}$$
$$C_5 = 24\ \mu\mu\text{F}$$
$$\omega_0 = 2{\cdot}442 \times 10^6 \text{ per sec.}$$
$$R = 2{\cdot}1 \text{ ohms}$$
$$L = 160\ \mu\text{H}$$

The value of C_0 is also required; it is calculated from L and ω_0 and found to be 1 047 $\mu\mu$F.

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It would be found that, with this value of C_0 and the values of C as read on the curves of Fig. 17, the calculated frequency curves would be 4 $\mu\mu$F to the right of the experimental curves. It is therefore necessary to alter C_0 slightly, so as to obtain agreement. An approximate formula for C_0 in terms of the experimental points of cut-off, C' and C'', is

$$(C_0/\beta) = \tfrac{1}{2}(C' + C'') + \tfrac{1}{2}C_3$$

The experimental cut-offs of the two curves under consideration are 1 043 and 1 040 $\mu\mu$F, and C_0, as calculated from the approximate formula, is 1 043 $\mu\mu$F. We shall, for convenience, adopt this value throughout the paper.

There is not much to choose between the two determinations of C_0. In the first case, the inductance L of the oscillatory circuit is required; it is not known, however, to much better than 5 parts in 1 000. In the second case, the final experimental figures for C' and C'' demand the estimation of the stray capacity between

TABLE 7.

Resonator Connected between Plate and Grid.

C	$\alpha - 1$	$f = 389\,000 +$	f (adjusted to fit the experimental value for large values of C) = 389 000 +
1	2	3	4
1 250	0·008	9·4	16·4
1 150	0·015	10	17
1 100	0·027	10·6	17·6
1 070	0·053	12·5	19·5
1 050	0·142	18	25
1 047	0·189	21	28
1 045	0·243	24	31
1 044	0·283	27	34

the leads connecting the plate and the filament to the condenser terminals. This stray capacity was taken as 6 $\mu\mu$F, which is perhaps under-estimated. Perfect agreement would be obtained by making it 10 $\mu\mu$F. The alternative, which has been adopted here for convenience, is to suppose L slightly larger, namely, 160·6 μH; and C_0 correspondingly smaller, namely, 1 043 $\mu\mu$F instead of 1 047 $\mu\mu$F.

Calculations have been performed using the values already adopted for the various constants, and the results are given in Table 7. Col. 3 gives the calculated frequency. This frequency is, for large values of the capacity, say 1 150 $\mu\mu$F, approximately 7 cycles per sec. below the experimental value. This constitutes an error of about 6 per cent in the value of $f - f_0$. Greater accuracy cannot, however, be expected, since repeated experimental work has proved that, even in the simpler case of a quartz resonator, the formula

$$f = f_0\left\{1 + \tfrac{1}{2}(K/K_t)\right\}$$

does not account for the change of frequency which actually accompanies a change in air-gap.

The values in col. 4 of Table 7 were obtained by adding 7 to the figures in col. 3; this is equivalent to making the theoretical curve fit the experimental curve for large values of C, that is, for points where the rate of change of frequency with respect to C is small. A

TABLE 8.

Resonator Connected between Grid and Filament.

C	$1 - \alpha$	$f = 389\,000 +$	f (adjusted to fit the experimental value for small C) = 389 000 +
850	0·0014	8·8	12·1
950	0·0029	8·6	11·9
1 000	0·0065	8·4	11·7
1 020	0·013	8·0	11·3
1 030	0·025	7·1	10·4
1 035	0·049	5·5	8·8
1 038	0·113	1·0	4·3
1 039	0·200	− 5·5	− 2·2
1 039·5	0·325	− 15	− 12

fair test of the theory is then to see how far the theoretical curve fits the experimental curve at the other points, in particular at points where the rate of variation of frequency is large. The two curves are shown in Fig. 18; the theoretical curve bends more sharply than the experimental curve. If the value of C_2 were made slightly larger than 6·4 $\mu\mu$F the two curves could be made to fit everywhere. In fact, the effective value of C_2 is larger than 6·4, because the grid leak introduces from 0·1 to 0·2 $\mu\mu$F, and there must be some capacity between the filament lead and the lead joining the resonator to the grid of the valve.

Similar calculations were performed for the grid-to-filament type of connection, and the calculations are given in Table 8.

TABLE 9.

Resonator Connected between Grid and Filament.

C	$1 - \alpha$	= 389 000 +	f (adjusted to fit the experimental value for small C) = 389 000 +
850	0·003	8·7	12·1
950	0·006	8·5	11·9
1 000	0·014	8	11·4
1 020	0·027	7	10·4
1 030	0·052	5·2	8·6
1 035	0·095	2·3	5·7
1 038	0·190	− 1·0	− 2·4
1 039	0·285	− 12	− 9

The value of C_3 was taken as 2·5 $\mu\mu$F as measured on the Schering bridge. The theoretical curve and the experimental curve are shown in Fig. 18. Here, again, the theoretical curve has a sharper bend than the experimental curve. The points marked on or near the experimental curve were calculated for a value of C_3

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equal to 3·7 $\mu\mu$F, and a value of C_0 equal to 1 044·6; the values are given in Table 9.

It is thus seen that a larger value of C_3, accompanied by a slight adjustment of C_0, brings the theoretical and experimental values into almost complete agreement. The small adjustment in C_0 is necessary to make the capacity C, at which the oscillations cease, equal in both cases. The alteration of C_3 is in the right direction because, in addition to the capacity between grid and plate, measured on the bridge, there is some capacity between the lead connecting the plate to the condenser, and the lead connecting the grid to the resonator. This may well have been 1 $\mu\mu$F.

between the two resonator leads may be altered by a considerable amount, say 1 $\mu\mu$F. This alters the value of K_3 and therefore of C_5. Suppose in one case C_5 was 24 $\mu\mu$F and in the other case 23 $\mu\mu$F. The corresponding difference in frequency, obtained from equation (49), is just 3 cycles per sec., which is about equal to the difference found experimentally.

Curves I and II, Fig. 17, exhibit a decrease of $|df/dC|$ with decreasing values of C in the neighbourhood of the cut-off point. This decrease is not in accordance with the approximate formulæ worked out in the paper; it is nevertheless a real effect, and not an experimental error. The reason why it is apparent with

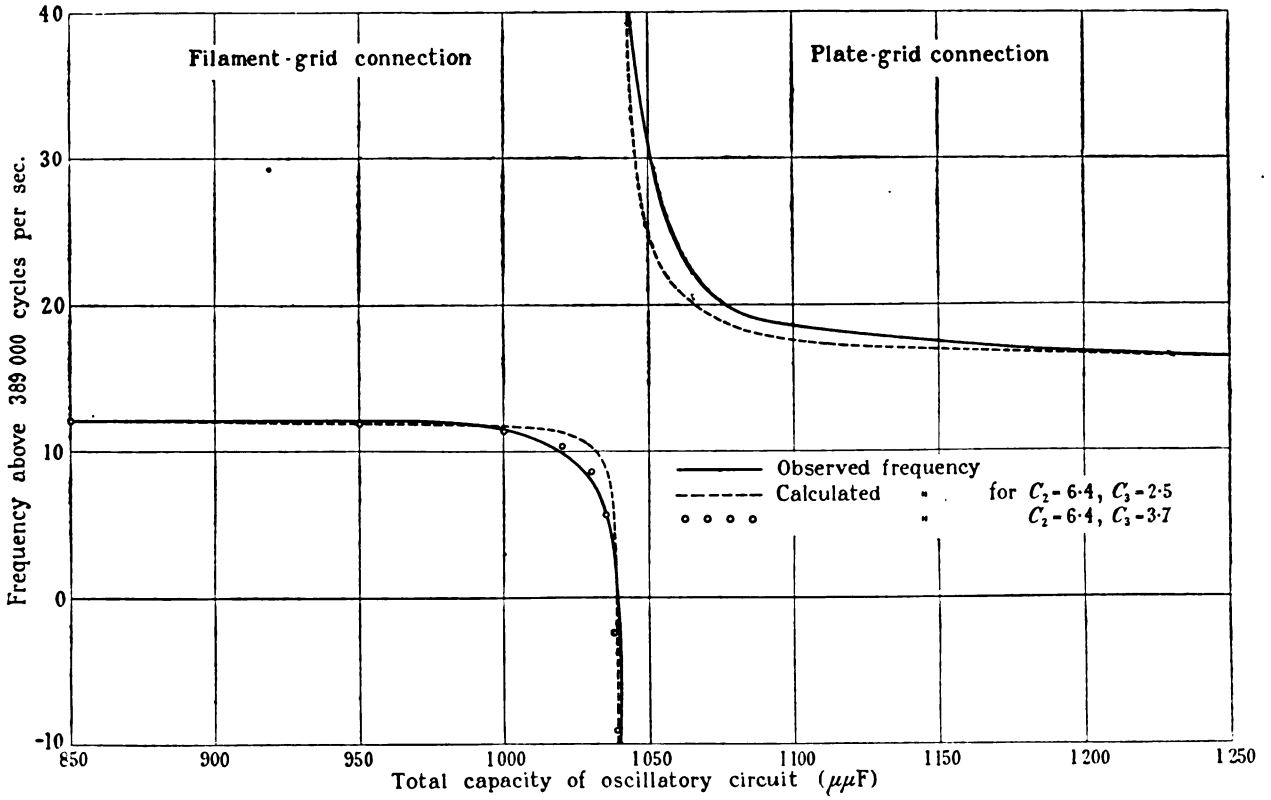

FIG. 18.—Comparison between calculated and observed frequencies for various capacities.
Valve: D.E.R. L = 160 μH; R = 2·1 ohms; grid leak = 0·36 × 10^{-6} mho.

The frequency for large values of C in the plate-to-grid type of connection, and the frequency for small values of C in the grid-to-filament type of connection, should be almost equal to each other, and should be independent of the grid leaks. The asymptotic frequency is given by equation (49). Fig. 17 shows, however, that the asymptotic value varies by one or two cycles when the grid conductance is increased from $0{\cdot}4 \times 10^{-6}$ mho to 10 or 12×10^{-6} mho. This variation may be accounted for by the small capacity introduced by the grid leaks.

When it is desired to change from one type of connection to the other, the resonator lead is disconnected, say, from the plate, and connected to the filament; the other resonator lead is left permanently connected to the grid. In this process of changing over, the capacity

large grid conductances and not with small ones may be due to the fact that the range of capacity over which this effect occurs becomes smaller and smaller as the grid conductance is decreased.

This divergence between predicted results and experimental results might be due to the variation of the valve plate conductance g with plate current. The decrease in conductance g would tend to reduce the increase in α, as can be seen from equation (37). It is quite as possible, however, that the cause lies in the approximations made when establishing the formulæ.

Relation between Frequency and Inter-electrode Capacities.—It has been shown on pages 47 and 52 that the frequency decreased with C_2 and that the modulus $|dn/dC_2|$ of the rate of change also decreased with C_2

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for both types of connection. It has also been shown that the frequency decreases with C_3 for both types of connection.

In order to confirm this prediction, a small variable

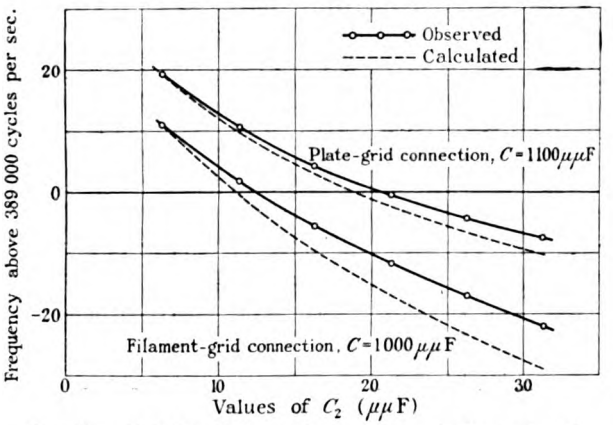

FIG. 19.—Relation between frequency and filament-grid capacity C_2.

Valve: D.E.R. $L = 160\ \mu H$; $R = 2{\cdot}1$ ohms; grid leak $= 0{\cdot}8 \times 10^{-6}$ mho; $C_3 = 2{\cdot}4\ \mu\mu F$.

condenser, of capacity 4 to 24 $\mu\mu$F, was placed in parallel with either C_2 or C_3, and the frequency was measured for various settings of that condenser.

The theory is thus being tested for frequency differences from a point on the curve, not for frequencies

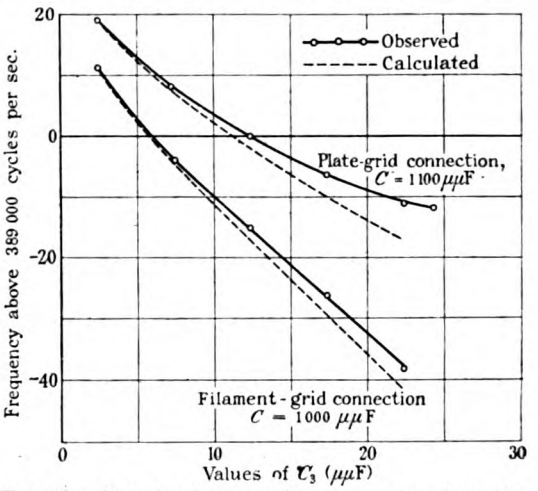

FIG. 20.—Relation between frequency and plate-grid capacity C_3.

Valve: D.E.R. $L = 160\ \mu H$; $R = 2{\cdot}1$ ohms; grid leak $= 0{\cdot}8 \times 10^{-6}$ mho; $C_2 = 6{\cdot}4\ \mu\mu F$.

above f_0. The reason for this has already been given on page 47.

Equations (47) and (70) were used to calculate Δf,

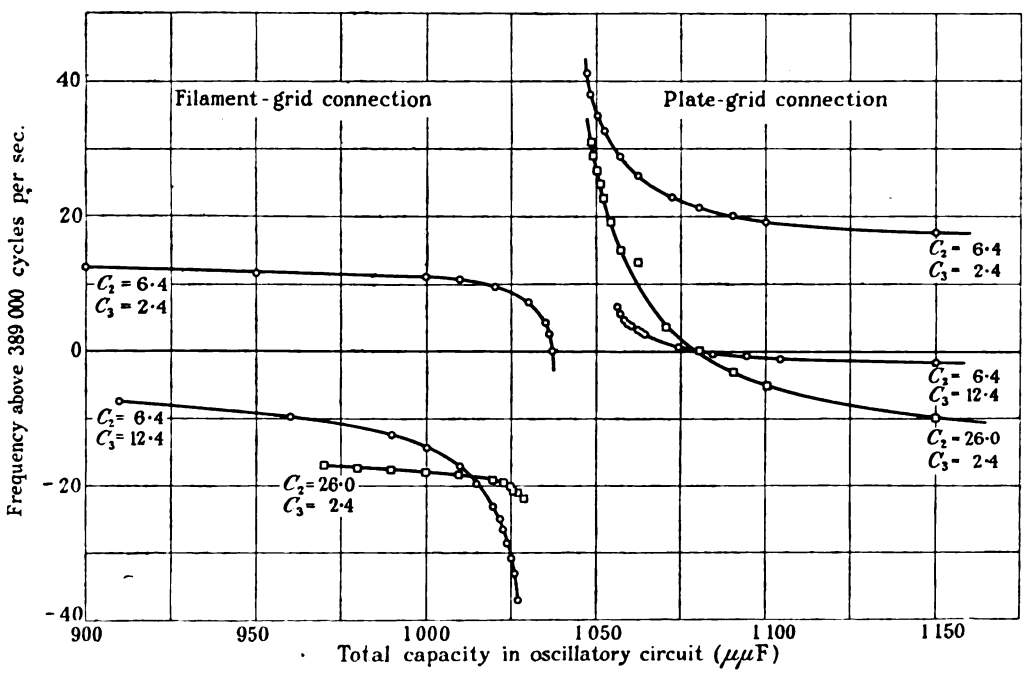

FIG. 21.—Relation between frequency and inter-electrode capacities.

Valve: D.E.R. $L = 160\ \mu H$; $R = 2{\cdot}1$ ohms; grid leak $= 0{\cdot}8 \times 10^{-6}$ mho.

The experimental curves are shown in Figs. 19 and 20, and their shape confirms the theoretical conclusions. The theoretical curves were made to fit the experimental curves at one point only, namely, the point for hich $C_2 = 6{\cdot}4$ and $C_3 = 2{\cdot}4$.

so that the effect of the resistance R and the inter-electrode conductances on the frequency differences were neglected. The values taken for C_2, C_3 and K_3 were obtained by separate measurements, entirely independent of the experiments on the oscillator. The theoretical

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curves could be made to fit the experimental curves much more nearly by slightly altering those values, but it has not been deemed necessary to do this here.

Fig. 21 shows the frequency variations with respect to C for various values of C_2 and C_3. The shape of the curves agrees with the deductions made on theoretical grounds.

As an example, consider the plate-to-grid type of connection, for which df/dC is given by equation (48). It is always negative and increases numerically with C_2. Hence, for given values of C_3 and K_3, the slope of the curve is greater when C_2 is large than it is when C_2 is small. This point is well borne out by the experimental curves.

of R, other things being the same, and the frequency is calculated by the formula

$$f = f_0 + \frac{1}{2} f_0 \frac{K}{K_t} \frac{\alpha K_2 + C_5}{\alpha(K_1 K_2 / K_t) + C_5}$$

For any given value of the total capacity C, the amplitude of oscillations decreases considerably with R; it is to be expected that the effective anode conductance g alters considerably in the process. Even then the formula could not account for the large variations of frequency taking place.

As an example, with $g_2 = 0{\cdot}5 \times 10^{-6}$, $g_3 = 0{\cdot}05 \times 10^{-6}$, $g_4 = 0{\cdot}8 \times 10^{-6}$, and $g = 25 \times 10^{-6}$, when R increases from 2 to 12 ohms, the frequency for $C = 1\,100$ $\mu\mu$F

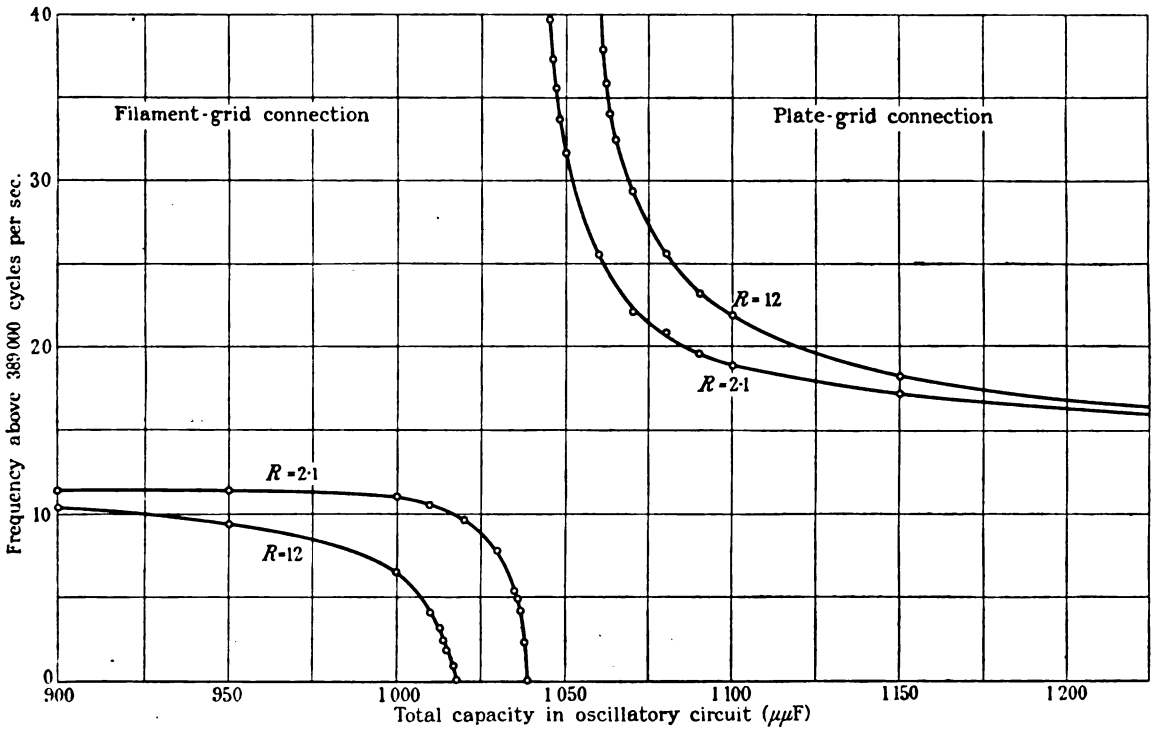

FIG. 22.—Relation between frequency and resistance of oscillatory circuit.
Valve: D.E.R. $L = 160$ μH; grid leak $= 0{\cdot}8 \times 10^{-6}$ mho.

Relation between Frequency and Resistance of the Oscillatory Circuit.—Curves of frequency against total capacity C were taken for various values of the resistance R of the oscillatory circuit. In order to alter the resistance of that circuit, the link closing the circuit was replaced by non-inductive resistances. The resistances were originally made for the measurement of the effective resistance of circuits at radio frequencies by the resistance-variation method, and their effect on the inductance and on the capacity of the circuit is known to be negligible.

The curves obtained are shown in Fig. 22. The variations are in the direction predicted on page 47, and when the oscillatory circuit is considerably out of tune the frequency is practically independent of the resistance. The actual variations in frequency are, however, larger than the figures to be expected from equations (37) and (68). These formulæ give α in terms

increases by 0·7 cycle per sec., according to the formula. If at the same time g is supposed to have increased from 25×10^{-6} to 40×10^{-6}, the total calculated increase in frequency is 1 cycle per sec.

The experimental figure is 3 cycles per sec. Part of the difference may, of course, be due to the uncertainty in the values of the inter-electrode conductances g_2 and g_3.

It must be remembered that equation (37) for α is only an approximation, which supposes that $HE - GF = 0$. This is equivalent to writing $\beta_2 = 0$ [see equations (29) and (36)]. But when a constant amplitude is being maintained, we have, from equation (19), $\beta_2 + SK = 0$. The same expression also shows that, near the cut-off point, the terms H, W and F become important, especially when R is large. The error in the approximation is then larger.

Relation between Amplitude of Oscillations and Capacity

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of Oscillatory Circuit.—A number of curves were obtained for various values of added grid-to-filament conductance, ranging between zero and 12×10^{-6} mho. These are all reproduced in Fig. 23.

The shape of the curves is analogous to that which was predicted from theoretical considerations. The amplitude is small when the oscillatory circuit is considerably out of tune. As the capacity C is varied towards C_0, the amplitude increases more and more rapidly and reaches a maximum at a value of C which is not very different from C_0. From this maximum it decreases very rapidly and then drops suddenly to zero.

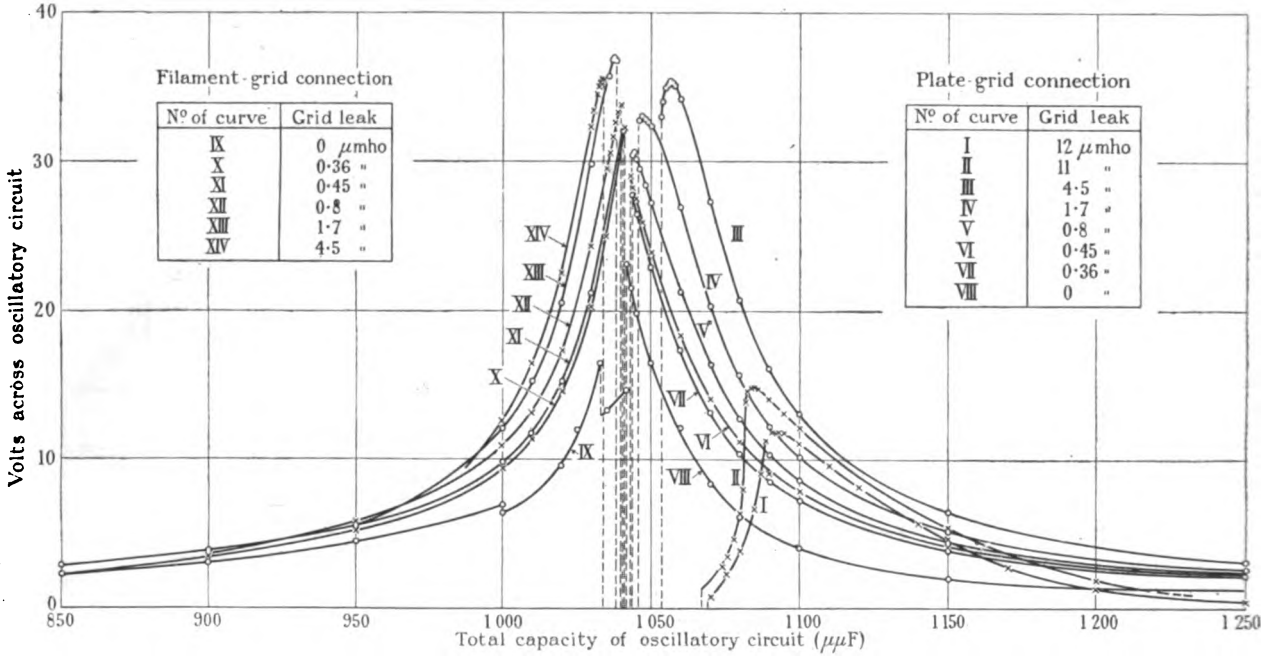

FIG. 23.—Relation between amplitude and capacity of oscillatory circuit.
Valve: D.E.R. $L = 160$ μH; $R = 2{\cdot}1$ ohms.

The value of C at cut-off is very close to the value of C for maximum amplitude.

It is seen from the curves that the largest maximum amplitude is given in one case by an added grid conductance of $4{\cdot}5 \times 10^{-6}$ mho, and in the other case by an added grid conductance of $1{\cdot}7 \times 10^{-6}$ mho. The formulæ developed do not explain this; on the contrary, the theoretical curves (Fig. 6) indicate that an increase in the grid conductance g_4 is accompanied by a decrease in maximum amplitude. It has already been mentioned, however, that other considerations arise; for example, the grid leak may tend to keep the mean potential of the grid at the most favourable place on the characteristic curve of the valve.

In Fig. 23 the amplitude curve for the filament-to-grid connection without any grid leak shows discontinuities at a few points. The reason for this is that the second mode of vibration already mentioned caused oscillations simultaneous with those under observation, for certain values of the capacity C.

On all other points there is fair agreement between theory and experiment. The plate-to-grid curves do not fall off as rapidly as the grid-to-filament curves, the reason being that C_2 was larger than C_3. The effect of the grid leaks on the cut-off point is in the direction predicted by theory, and the order of magnitude is also the same.

Oscillations could not be started in the filament-to-grid case when the added grid conductance was 11×10^{-6} mho or larger.

The maximum point and the cut-off point are nearer together in practice than was to be expected from the theoretical curves of Fig. 6, especially in the case when the added grid conductance was small, namely $0{\cdot}8 \times 10^{-6}$ mho. In fact, the points were so near together in practice that it was impossible to distinguish them. The reason for this divergence seems to be that the value of 400 ohms taken for S is too large. It was obtained by measurements on the quartz when used as a resonator. It is conceivable that, in an oscillator, the piece of quartz will vibrate itself more readily into the position giving minimum damping.

TABLE 10.

Type of connection	Added grid conductance	Range	
		Calculated	Observed
	mho × 10⁻⁶		
Plate-grid ..	0·8	1 540 to 1 053	1 800 to 1 044
Plate-grid ..	4·5	1 290 to 1 061	1 350 to 1 053
Filament-grid	0·8	850 to 1 036	800 to 1 040
Filament-grid	4·5	940 to 1 034	880 to 1 034

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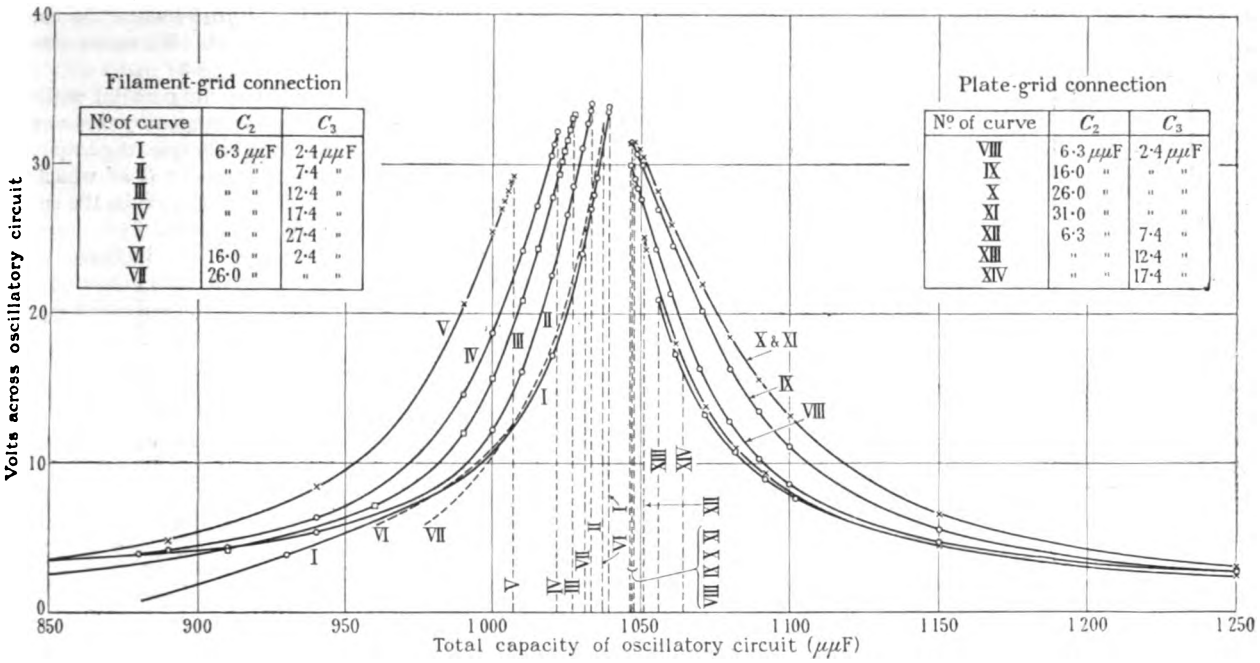
Filament-grid connection
Nº of curve | C_2 | C_3
I | 6·3 μμF | 2·4 μμF
II | " " | 7·4 "
III | " " | 12·4 "
IV | " " | 17·4 "
V | " " | 27·4 "
VI | 16·0 " | 2·4 "
VII | 26·0 " | " "
Plate-grid connection
Nº of curve | C_2 | C_3
VIII | 6·3 μμF | 2·4 μμF
IX | 16·0 " | " "
X | 26·0 " | " "
XI | 31·0 " | " "
XII | 6·3 " | 7·4 "
XIII | " " | 12·4 "
XIV | " " | 17·4 "
Volts across oscillatory circuit
Total capacity of oscillatory circuit (μμF)

FIG. 24.—Relation between amplitude and inter-electrode capacities.

Valve: D.E.R. $L = 160$ μH; grid leak $= 0{\cdot}8 \times 10^{-6}$ mho.

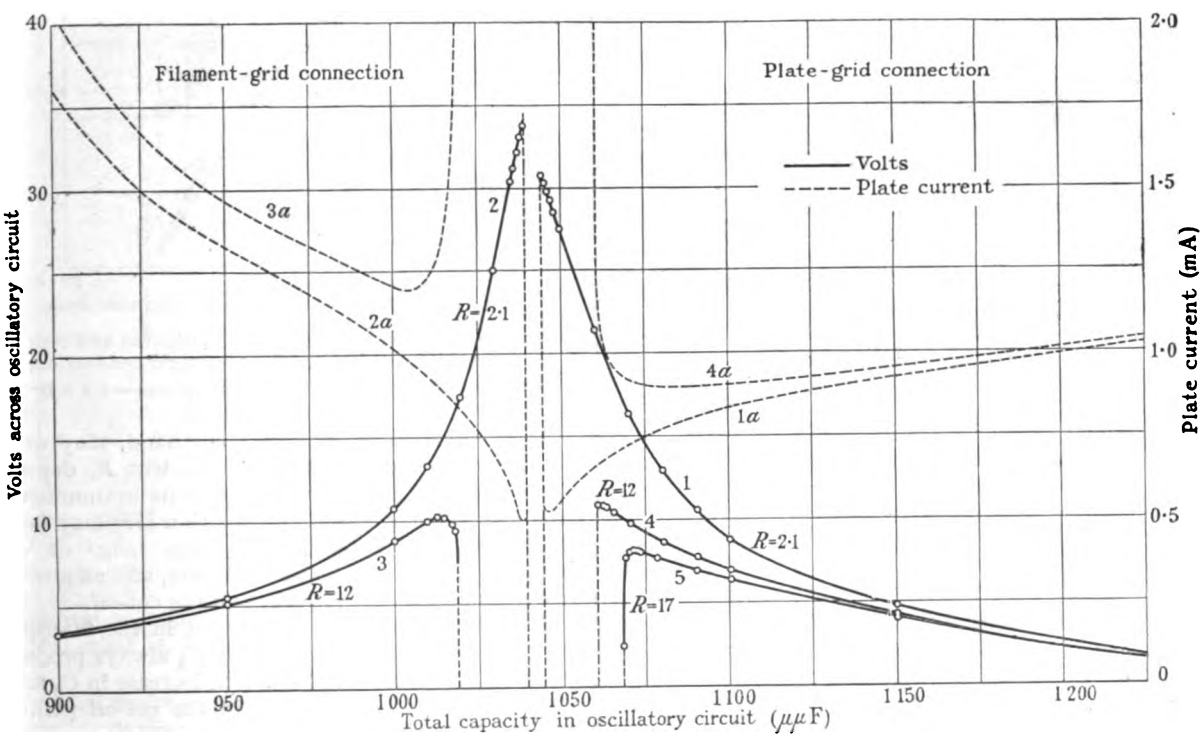

FIG. 25.—Relation between amplitude and resistance of oscillatory circuit.

Valve: D.E.R. $L = 160$ μH; grid leak $= 0{\cdot}8 \times 10^{-6}$ mho.

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The difference between the range of capacity over which oscillations were maintained might also be accounted for by this over-estimation of S. In Table 10 some figures are given to show the difference.

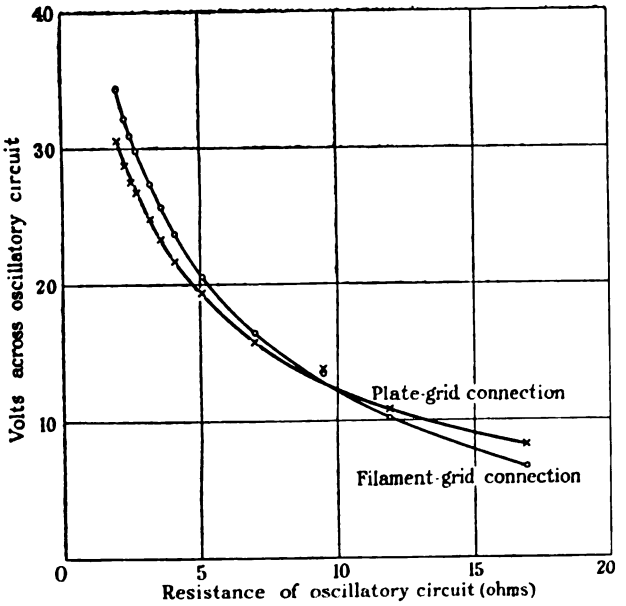

FIG. 26.—Relation between maximum amplitude and resistance of oscillatory circuit.
Valve: D.E.R. $L = 160\ \mu H$; grid leak $= 0{\cdot}8 \times 10^{-6}$ mho.

Relation between Amplitude of Oscillations and Inter-electrode Capacities.—The experimental results are embodied in the curves of Fig. 24. The variations are in accordance with the theory.

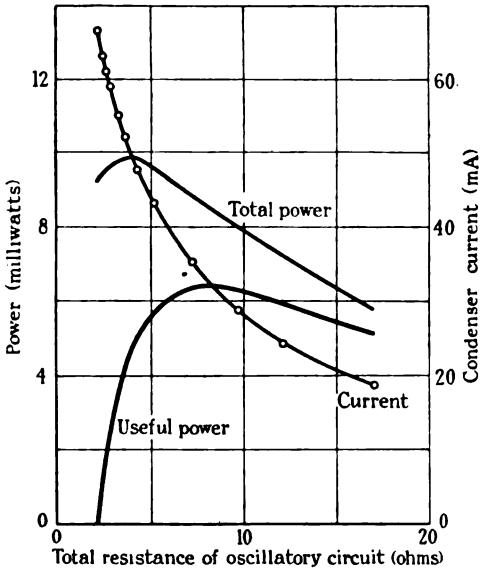

FIG. 27.—Relation between power dissipated and resistance of oscillatory circuit. Plate-grid connection.
Valve: D.E.R. $L = 160\ \mu H$; $R = 2{\cdot}1$ ohms; grid leak $= 0{\cdot}8 \times 10^{-6}$ mho.

In the case of the plate-to-grid connection, the amplitude increases at first as C_2 is increased. There is an optimum value of C_2 past which the amplitude would decrease. In the present case this optimum seems to be between 26 and 31 $\mu\mu F$, since the curves for those two capacities are identical. No measurements were made with a capacity exceeding 31 $\mu\mu F$.

An increase in C_3, the capacity in parallel with the quartz resonator, is always disadvantageous, as shown by curves XII, XIII and XIV. Not only does the amplitude decrease, but the value of the capacity C at which the cut-off occurs is increased. This variation in the cut-off point is readily explained by the theory.

In the plate-to-grid connection, an increase in C_3 or K_3 is equivalent to a decrease in E in equation (54). The value of x required to satisfy the equation $\phi_2 = 0$ [see equation (55)] is therefore increased.

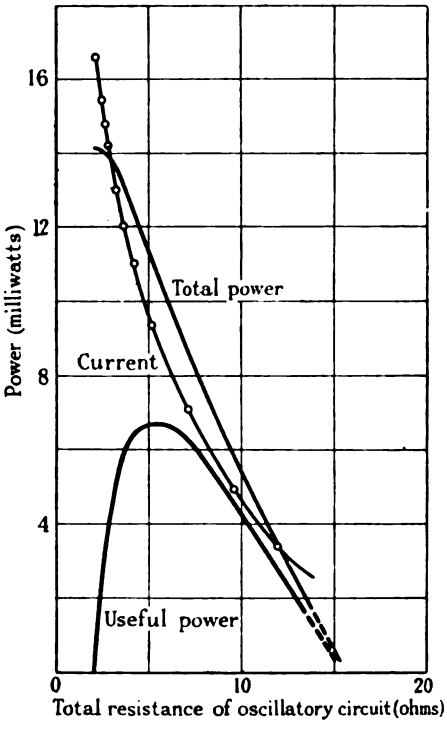

FIG. 28.—Relation between power dissipated and resistance of oscillatory circuit. Filament-grid connection.
Valve: D.E.R. $L = 160\ \mu H$; $R = 2{\cdot}1$ ohms; grid leak $= 0{\cdot}8 \times 10^{-6}$ mho.

An increase in C_2, on the other hand, may or may not mean an increase in the coefficient E, depending upon whether C_2 is below or above its optimum value. Moreover, the coefficients in equation (54) may be such that $|\partial\phi_2/\partial C_2|$ is small over a large range of values of C_2. This seems to be the case here, and an alteration in C_2 has no appreciable effect on the cut-off.

Similar considerations show that in the filament-to-grid case an increase in C_2 or in K_3 always produces a decrease in the cut-off point. An increase in C_3 may or may not produce a decrease in the cut-off point. It does so in the present case. It is to be noted that a decrease in the cut-off capacity is not necessarily accompanied by a decrease in the maximum amplitude.

Relation between Amplitude of Oscillations and Resistance of the Oscillatory Circuit.—Complete amplitude curves for

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different resistances are shown in Fig. 25. When the resistance of the oscillatory circuit is large, the maximum amplitude is small, and the values of C for maximum amplitude and for cut-off are further apart than is the case with a low-resistance oscillatory circuit.

When the oscillatory circuit is considerably out of tune the amplitude is approximately inversely proportional to the impedance of the oscillatory circuit, which is nearly equal to

$$\omega_0 L / \sqrt{\{\omega_0^2 R^2 C^2 + (1 - \omega_0^2 LC)^2\}}$$

This expression is practically independent of R when $1 - \omega_0^2 LC$, that is, $1 - (C/C_0)$ is as large as, say, 0·3 or 0·4.

The added resistance, just as the added grid conductances, produces an earlier cut-off. This is in perfect agreement with the theoretical curves (Fig. 6). Theory and experiment both show, moreover, that the other limiting capacity, given by equations (42) and (73), is not appreciably altered by addition of resistance; the range is therefore approximately the same, whatever be the resistance of the oscillatory circuit, whereas it is greatly reduced by addition of grid conductance, or by an increase in the damping S of the quartz.

The curves in Fig. 26 give the maximum amplitude for various values of the resistance R of the oscillatory circuit. They confirm the prediction that the amplitude is approximately inversely proportional to the resistance of the oscillatory circuit. The departure from the hyperbolic law is also in the direction predicted. This rule is not confined to quartz oscillators: it is perfectly general, and demonstrates the importance of designing low-resistance coils for valve oscillators.

When the resistance of the oscillatory circuit is large, the maximum amplitude for the plate-to-grid connection is larger than the maximum amplitude for the filament-to-grid connection. This was found to be the case for all the valves and all the inductances used. From the theoretical curves (Fig. 6) one might have expected just the reverse phenomenon. Here, again, the mean grid

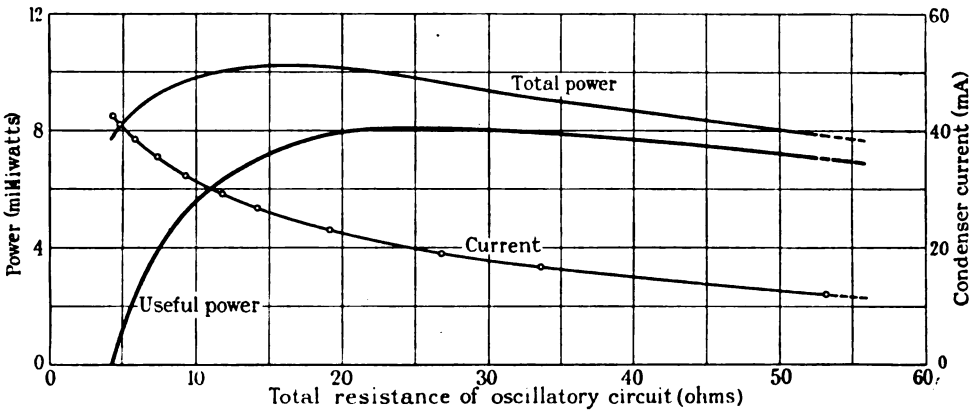

FIG. 29.—Relation between power dissipated and resistance of oscillatory circuit. Plate-grid connection.
Valve: D.E.R. $L = 325$ μH; $R = 4{\cdot}3$ ohms; grid leak = $0{\cdot}8 \times 10^{-6}$ mho.

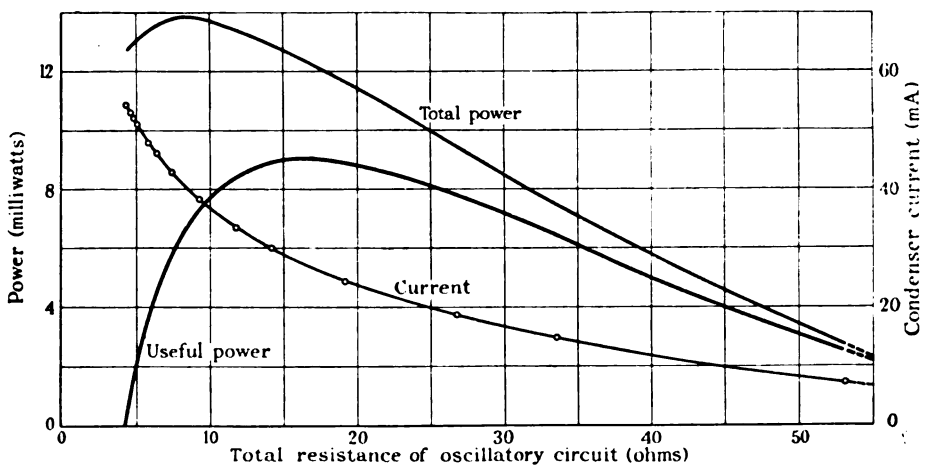

FIG. 30.—Relation between power dissipated and resistance of oscillatory circuit. Filament-grid connection.
Valve: D.E.R. $L = 325$ μH; $R = 4{\cdot}3$ ohms; grid leak = $0{\cdot}8 \times 10^{-6}$ mho.

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potential is probably governed by other factors besides the grid leak and the amplitude.

The point could easily be studied further by eliminating x between the equations

$$\phi = 0 \quad \text{[see equation (54)]}$$
and $$(\partial\phi/\partial x) = 0 \quad \text{[see equation (57)]}$$

solving for $1/g$ and assigning various values to R for both types of connections.

Relation between Power Dissipated and Resistance of Oscillatory Circuit.—In all the measurements described below, the condenser current measured was the maximum current with respect to the capacity C. It corresponds to the maximum ordinates of the curves of Fig. 25. It is not always possible or advisable to work at maximum amplitude, because the maximum amplitude is so very

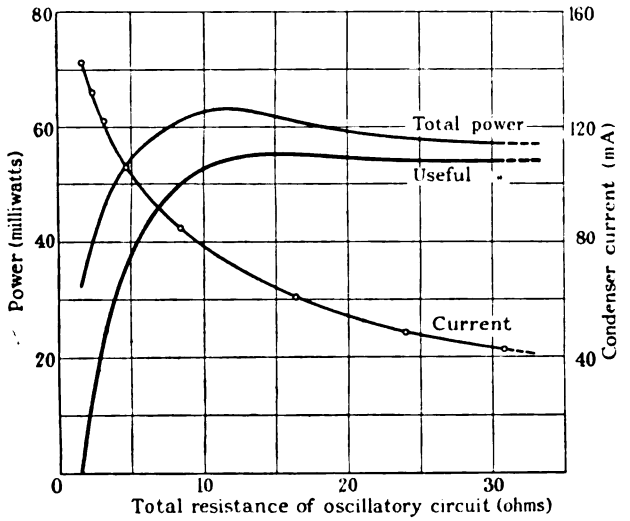

FIG. 31.—Relation between power dissipated and resistance of oscillatory circuit. Plate-grid connection.
Valve: D.E.P. 610. $L = 160$ μH; $R = 1{\cdot}57$ ohms; grid leak $= 0{\cdot}8 \times 10^{-6}$ mho.

near to the cut-off position, especially when the resistance of the oscillatory circuit is low. In cases where a large power is taken from the oscillator, it is possible to work at maximum amplitude, because, as can be seen from curves III and V, Fig. 25, the maximum is flatter, and not too near to the cut-off position. Also, since curves III and V lie wholly within curves II and I, Fig. 25, on sudden removal of the load the oscillator will go on working at a stable point of curves II or I.

For all the power measurements below, the added grid conductance was $0{\cdot}8 \times 10^{-6}$ mho.

For the sake of comparison, measurements were made with both types of connection, for the two circuits the constants of which are given on page 57 and the two valves the constants of which are given on page 56. The same plate voltage, viz. 100 volts, was used throughout.

The useful power is the product of the square of the current in the condenser C and of the added resistance R'. Under normal working conditions, R' is an equivalent series resistance, due to a coupled circuit or some other form of load. The total power is the product of $R + R'$ and the square of the condenser current. It is a more instructive figure than the former because it is dependent only on the sum of the resistances R and R', and not on R. It represents the available power in a circuit of zero inherent resistance.

If the amplitude were strictly inversely proportional to the resistance, the total power dissipated would vary in an exactly similar manner to the amplitude. A glance at Figs. 27 to 34 shows that this is not so. The total power would in practice be very small for a very small total resistance; as the total resistance is increased it reaches a maximum, and then decreases gradually. For large values of the total resistance, its variation becomes very similar to the variation of amplitude; for larger values still, no oscillations can be started.

The curves show in striking manner the difference

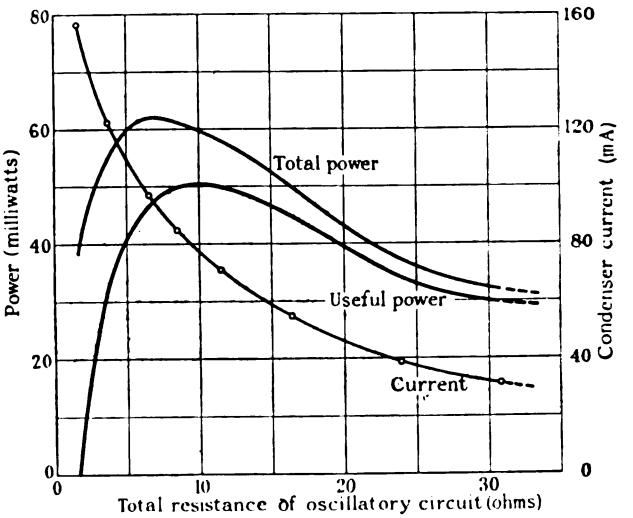

FIG. 32.—Relation between power dissipated and resistance of oscillatory circuit. Filament-grid connection.
Valve: D.E.P. 610. $L = 160$ μH; $R = 1{\cdot}57$ ohms; grid leak $= 0{\cdot}8 \times 10^{-6}$ mho.

between the plate-to-grid and filament-to-grid types of connection. In the filament-to-grid type, the maximum total power is in some cases higher than the corresponding power for the plate-to-grid connection. In all cases, however, the curve is sharper and the maximum power corresponds to a smaller total resistance than with the plate-to-grid connection. The plate-to-grid connection is therefore preferable, because it is advantageous to dispose of approximately maximum power over as wide a range of current as possible. Another advantage of the plate-to-grid connection is that subsidiary resonances of the quartz plate, in the neighbourhood of the main resonance frequency, do not produce secondary oscillations as readily as they seem to do with the filament-to-grid connection. On the other hand, for a given amplitude of plate voltage, the voltage across the quartz resonator is higher in the plate-to-grid type of connection than it is in the filament-to-grid connection. The swing E of grid voltage is, in general, smaller than half the amplitude V of the plate voltage. In consequence

$$|V - E| \geqslant |V| - |E| > |E|$$

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The high-conductance valve gives, other things being equal, much more power than the low-conductance valve. Moreover, oscillations are maintained for a much larger range of the resistance $R + R'$ and the curve of total power dissipated against total resistance of the oscillatory circuit is much flatter with the high-

of L to C_0. In the present case, with the D.E.R. valve, the ratio $L/C = 0{\cdot}6 \times 10^6$ is more advantageous for most purposes than the ratio $0{\cdot}15 \times 10^6$. The optimum ratio of L to C_0 depends upon the valve and the requirements of the user. Thus, with the D.E.P. valve the maximum power dissipated is approximately the same

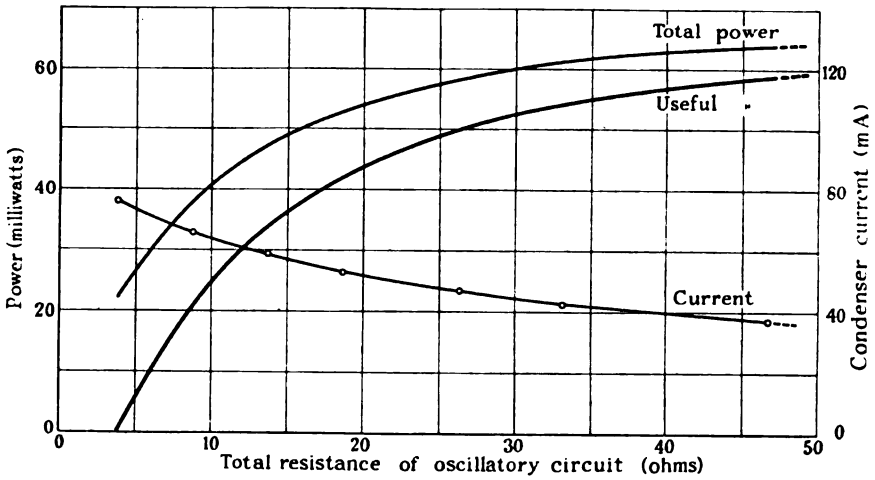

FIG. 33.—Relation between power dissipated and resistance of oscillatory circuit. Plate-grid connection.
Valve: D.E.P. 610. $L = 325\ \mu H$; $R = 3{\cdot}8$ ohms; grid leak $= 0{\cdot}8 \times 10^{-6}$ mho.

conductance valve. This provides greater flexibility for the use of the maximum output, and is a distinct advantage.

The amplitude of the voltage across the condenser increases when the inductance L is made larger, and

with the 160-μH and 325-μH coils. Suppose it is required to pass current through a resistance of 10 ohms and obtain as much power as possible. Figs. 31 to 34 show that the 160-μH coil is much more suitable than the 325-μH coil.

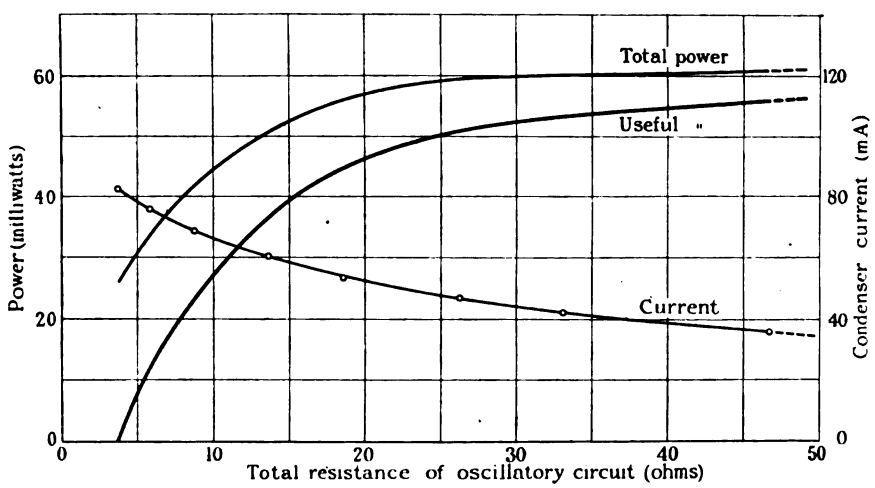

FIG. 34.—Relation between power dissipated and resistance of oscillatory circuit. Filament-grid connection.
Valve: D.E.P. 610. $L = 325\ \mu H$; $R = 3{\cdot}8$ ohms; grid leak $= 0{\cdot}8 \times 10^{-6}$ mho.

the capacity C correspondingly smaller. The increase in amplitude for a given resistance is not, however, in the proportion of the square of the inductance.

The maximum is not very different for the two inductances, but oscillations are maintained for a much larger range of total resistance, and the power curves are much flatter, with the high inductance. It is, therefore, advisable to work with a fairly large ratio

When the power is dissipated in a circuit of inductance L_1 and resistance R_1 coupled to the coil L the above considerations are not so important, because a new variable is provided by the mutual inductance M. The equivalent resistance $R + R'$ is equal to $R + \left\{(\omega^2 M^2 R_1)/(R_1^2 + \omega^2 L_1^2)\right\}$ and, with a given value of R_1, can be altered by altering M until the condition for maximum power is reached.

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In the case of Figs. 27 and 28, oscillations could not be started with resistances of 24 ohms and above. In the other six cases oscillations could have been produced with resistances even larger than the largest used. In the case of Fig. 33, the maximum power was not reached although the total resistance was nearly 50 ohms.

The modifications of the Pierce circuits with a view to obtaining large powers from a crystal have not been included in this investigation. Circuits for this purpose have already been devised by others. Particular mention should be made of the work of J. M. Miller and A. Crossley,* who use, among other improvements, a choke coil and grid battery instead of a grid leak; the potential of the grid is thus kept negative over the greater part of the cycle, and the loss due to grid current is considerably reduced.

(3) CONCLUSION.

In the first part of the paper the behaviour of the valve-maintained quartz oscillator has been studied theoretically. The only circuits considered were the two variations of the Pierce circuit, namely:

(i) The plate-to-grid connection, in which the quartz resonator is connected between the plate and the grid; and

(ii) The filament-to-grid connection, in which the resonator is connected between the filament and the grid.

In both cases there was an oscillatory circuit between the plate and the filament.

The quartz oscillator has been represented by a network of admittances, and an equation obtained for the oscillations in the plate circuit. An approximate solution of this equation has been obtained, and expressions have been developed for the frequency and the amplitude of the oscillations in the plate circuit. From these expressions it could be deduced that:

(i) With the plate-to-grid connection:

(*a*) When the capacity C in the oscillatory plate circuit is several times as large as the value C' which would make the natural frequency of the oscillatory circuit equal to the frequency of the quartz resonator, no oscillations take place. As the capacity C is reduced gradually, a value is reached at which oscillations start at a very small amplitude. Further decrease of C is accompanied by increase in the amplitude, until the amplitude reaches a maximum at a value of C which is in general less than 1 per cent above C'. A further small decrease in C produces a rapid decrease in amplitude, until a value is reached at which the oscillations stop suddenly. This value also is greater than C'.

(*b*) The resistance of the oscillatory circuit, the grid leak, and, in general, any resistances or leaks in the circuit, tend to decrease the maximum amplitude of oscillations.

(*c*) A certain amount of capacity, either in the valve or outside, between the grid and the filament, is essential to the maintenance of oscillations. On the other hand, large values of this capacity will make the amplitude small. There is an optimum value which gives maximum power.

(*d*) Any capacity in parallel with the quartz resonator, whether inter-electrode capacity between the plate and the grid, stray capacity between the electrodes of the resonator, or capacity between the leads connecting the resonator to the valve, is always disadvantageous and reduces the power.

(*e*) A large plate conductance and a large amplification factor are beneficial.

(*f*) Any variation of frequency produced by changing the air-gap, the valve or the circuits is in all cases a small fraction only of the response frequency of the resonator when the air-gap is zero, i.e. when the electrodes of the resonator are in contact with the quartz crystal.

(*g*) When the capacity C is gradually decreased from a large value, the frequency increases slowly at first, then more and more rapidly, until the sudden cut-off of the oscillations.

(*h*) The greater the conductance of the grid leak, the greater the frequency.

(*j*) The variation of frequency for a given change of C is greater with large grid conductances than it is with small grid conductances.

(ii) With the filament-to-grid connection:

(*a*) When the capacity C in the plate circuit is zero, oscillations are weak or do not take place at all. As the capacity is gradually increased the amplitude of oscillations increases gradually, and reaches a maximum at a value of C which is, in general, less than 1 per cent below C'. A further small increase of C produces a rapid decrease in amplitude, until a value of C, also smaller than C', is reached, at which the oscillations stop suddenly.

(*b*) The resistance of the oscillatory circuit, the grid leak, and, in general, any resistances or leaks in the circuit, tend to decrease the maximum amplitude of oscillations.

(*c*) A certain amount of capacity, either in the valve or outside, between the grid and the plate, is essential to the maintenance of oscillations, but when this capacity is very large no oscillations take place. There is an optimum value of this capacity, which gives maximum power.

(*d*) Any capacity in parallel with the resonator, either in the valve or outside, reduces the power.

* See Reference (8).

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(*e*) The frequency is always lower than the frequency in the plate-to-grid connection, but it is never as low as the response frequency of the resonator.

(*f*) Any variation of frequency produced by changing the air-gap, the valve or the circuits is in all cases a small fraction only of the response frequency of the resonator for zero air-gap.

(*g*) When the capacity C is gradually increased from a small value, the frequency decreases slowly at first, then more and more rapidly, until the sudden cut-off of the oscillations.

(*h*) The greater the conductance of the grid leak, the lower the frequency.

(*j*) The variation of frequency for a given change of C is greater with large grid conductances than it is with small grid conductances.

The experimental investigation has confirmed all the deductions made above, for each type of connection, with one exception. It has been found that addition of a grid leak of a suitable value may increase the amplitude of oscillations. The reason may be that the grid leak tends to keep the mean potential of the grid at the most favourable point on the characteristic curve of the valve.

This work was carried out under the auspices of the Radio Research Board, to whom thanks are due for permission to publish it. The author also wishes to thank Dr. D. W. Dye for his constant help and encouragement throughout the investigation.

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HIGH-FREQUENCY RESISTANCE MEASUREMENT BY THE USE OF A VARIABLE MUTUAL INDUCTANCE.*

By W. JACKSON, M.Sc., Student.

(*Paper first received 20th August, and in final form 21st October*, 1929.)

SUMMARY.

The paper deals with a method of high-frequency resistance measurement employing a variable mutual inductance. The sources and magnitude of the errors likely to occur are dealt with mathematically. The results of coil resistance measurements are given and compared with the calculated values of resistance and with the results obtained by the " resistance variation " method. The application to condenser resistance measurement is considered.

TABLE OF CONTENTS.

(1) Introduction.
(2) Theory of the method.
(3) Effect of direct induced voltage in the secondary coil.
(4) Effect of harmonics in the working current.
(5) Experimental results.
(6) Conclusions.
(7) Application to condenser resistance measurement.
Appendix.

(1) INTRODUCTION.

The subject of high-frequency resistance measurement has received considerable attention and it offers many difficulties. Of the several methods of coil resistance measurement which have been developed the " resistance variation " method has received most consistent application, mainly by virtue of its simplicity. The serious disadvantage of the method lies in the fact that the circuit resistance measured gives the combined resistance of the coil and the variable tuning condenser. If the coil resistance is to be accurately determined a condenser of known power factor is therefore a necessity. Since a variable condenser of known power factor at high frequency is not readily obtainable, the results are always liable to give a value of resistance in excess of the true coil resistance required.

The present paper deals with a method of high-frequency resistance measurement suggested by E. B. Moullin,† by which the coil resistance can be determined independently by use of a variable mutual inductance. A simple modification of the circuit connections provides a method of condenser resistance measurement. Although the procedure is different, this latter method closely resembles a bridge previously used by A. Campbell for measurements at acoustic frequencies and known as the " Campbell sifter." The method of coil resistance measurement is, however, quite distinct from that of the Campbell bridge, since the condenser employed is outside the measuring circuit and is used only for tuning purposes.

The early part of the paper consists of a theoretical discussion of the method of coil resistance measurement and of the sources of error which are liable to occur. The later part gives a description of the experimental work carried out, together with a statement of the results obtained, and closes with a brief consideration of the method of condenser resistance measurement.

(2) THEORY OF THE METHOD.

Consider the circuit shown in Fig. 1. L_1 represents a coil the resistance R_1 of which is to be measured. C is a variable condenser forming with L_1 a resonant

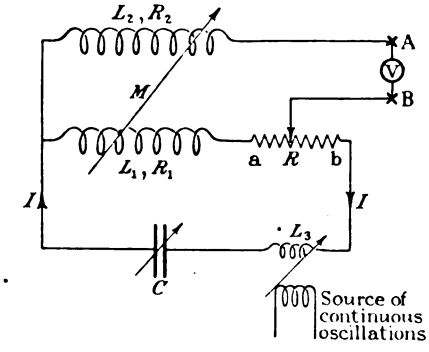

FIG. 1.

circuit coupled to a source of continuous oscillations, generating a current of frequency $\omega/(2\pi)$, by means of the small coupling coil L_3. L_2 is a coil whose coupling to L_1 can be varied and which forms with L_1 a variable mutual inductance M, in which the self and mutual inductive effects are in opposition. R is a potentiometer of resistance wire in series with the L_1CL_3 circuit, while V is an indicator for measuring the potential between the points A and B.

Let a current I of frequency $\omega/(2\pi)$ be caused to circulate in the circuit L_1CL_3 by tuning to the generator frequency. Provided that the indicator V has an infinite resistance, no current flows in the circuit L_2AB and the voltage between the points A and B may be written

$$V = \{R_1 + j(L_1 - M)\omega\}I$$

for the case in which the potentiometer tap is at " a."

* Reprinted from *Journal I.E.E.*, 1930, vol. 68, p. 296.
† " Radio-Frequency Measurements," p. 185.

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As M is varied the voltage V passes through a minimum value V_1 for $M = L_1$, when

$$V_1 = R_1 I \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

For this value of M the insertion between the points "a" and B of a pure resistance R increases the voltmeter reading to a value V_2

where $$V_2 = (R + R_1)I \quad . \quad . \quad (2)$$

Provided that I has been constant, equations (1) and (2) enable the unknown resistance R_1 to be determined in terms of V and R. It is preferable to vary R through a series of values and to plot a curve connecting V and the added resistance R. The curve ought to be a straight line, as in Fig. 2, having an intercept OR_1 along the R axis equal to the unknown resistance R_1. This reduces the error which a single reading might entail. If the current I is accurately known the value of R_1 may be checked as OV/I.

The application of this method to the determination of coil resistance at high frequency is, however, complicated by the self and mutual capacities which are

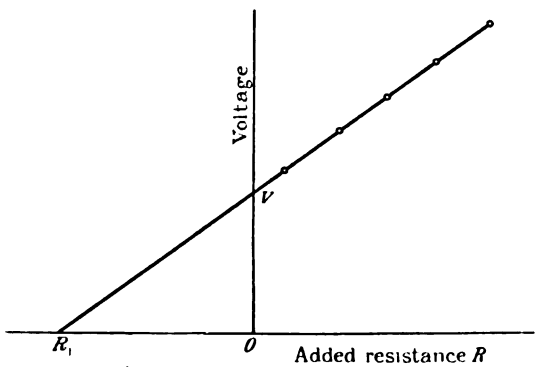

FIG. 2.

always associated with mutual inductances. As the frequency rises these capacities become of increasing importance and seriously modify the simple expression for the minimum voltage given by equation (1).

The self and mutual capacities of two coupled coils are distributed quantities the distribution of which varies with the nature of the individual coils and with their relative position. It is necessary, for simplicity, to represent these distributed quantities by lumped or concentrated capacities connected across the ends of the coils. While the current distribution in the various parts of the circuit to be considered may be seriously modified by earth capacity effects, it is difficult to allow for their effect since they are not so much a property of the circuit itself as of its position in relation to surrounding bodies and of its potential distribution. These effects, therefore, cannot be considered, but it is of importance to reduce them as much as possible by suitable circuit arrangement and careful earthing.

Representing the self-capacities of L_1 and L_2 by C_1 and C_2 and their mutual capacity by C_{12}, the circuit of Fig. 1 is given in diagrammatic form by Fig. 3.

The current I passing through the condenser C, and measured by the insertion of a thermo-ammeter in close proximity to C, is now the terminal current of L_1 and includes the current passing through C_1 and the other distributed capacities.

The required voltage between A and B, for a known input current I, may be derived as the vector sum of the voltages between AD and DB.

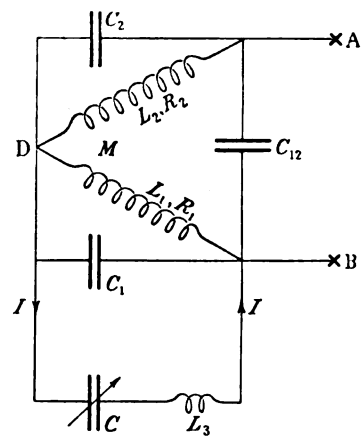

FIG. 3.

The potential difference existing between the terminals A and D of the secondary coil L_2 may be written

$$V'_S = (\sigma + jM'\omega)I \quad . \quad . \quad . \quad . \quad . \quad (3)$$

where M' is the effective mutual inductance between L_1 and L_2, including the effects of self and mutual capacity, and σ an impurity term for which these capacities are responsible.

In an analysis of the capacity effects in standard

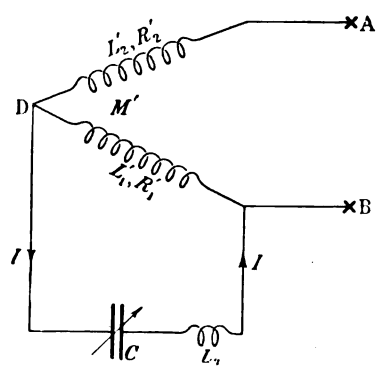

FIG. 4.

mutuals Butterworth * has shown that for the case in which the mutual effect opposes the self-inductance, as applies in the present investigation, the circuit of Fig. 3 may be replaced by the simple circuit of Fig. 4, where R'_1, R'_2, L'_1 and L'_2 represent the effective resistances and inductances of the primary and secondary coils respectively.

* *Proceedings of the Physical Society of London*, 1921, vol. 33, p. 312.

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R_1' and L_1' are given by the expressions

$$R_1' = R_1\left[1 + 2\omega^2\{C_1L_1 + C_{12}(L_1 - M)\}\right]$$

$$L_1' = \left[L_1 + \omega^2\{C_1L_1^2 + C_2M^2 + C_{12}(L_1 - M)^2\} - (C_1 + C_{12})R_1^2\right]$$

The effective mutual inductance M' and the impurity σ may be expressed thus:—

$$M' = M + C_{12}R_1R_2 + \omega^2\{C_1L_1M + C_2L_2M - C_{12}(L_1 - M)(L_2 - M)\}$$

$$\sigma = \omega^2\left[C_1R_1M + C_2R_2M - C_{12}\{R_1(L_2 - M) + R_2(L_1 - M)\}\right]$$

The general expression for the potential difference between the points A and B, due to the terminal current I, is therefore

$$V' = I\left[R_1\{1 + 2\omega^2\{C_1L_1 + C_{12}(L_1 - M)\}\} - \omega^2\{C_1R_1M + C_2R_2M - C_{12}\{R_1(L_2 - M) + R_2(L_1 - M)\}\}\right] + j\omega I\left[L_1 + \omega^2\{C_1L_1^2 + C_2M^2 + C_{12}(L_1 - M)^2\} - (C_1 + C_{12})R_1^2 - M - C_{12}R_1R_2 - \omega^2\{C_1L_1M + C_2L_2M - C_{12}(L_1 - M)(L_2 - M)\}\right] \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (4)$$

$$= I(a + jb) \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad .

In order for a second harmonic to double the error due to self and mutual capacity, the ratio N/I would have to be $1/\sqrt{2}$ and for the nth harmonic $1/\sqrt{(n^2-2)}$.

From the above it would appear that the error due to harmonics is not likely to be serious in any case in which the circuit L_1CL_3 is tuned to the fundamental frequency of the oscillator.

(5) Experimental Results.

It is evident that unless care is taken in the construction of the mutual inductance, having as primary the coil of which the resistance is required, serious errors may be incurred. The provision of a suitable mutual inductance is a difficult matter, especially as very fine variation is necessary in order that there may be no difficulty in adjusting for minimum voltage between A and B (Fig. 1). The mutual inductance used consisted of two cylindrical single-layer coils wound on paxolin formers mounted coaxially end to end, the primary coil L_1 being fixed and the secondary L_2 movable. A coarse and a vernier adjustment were employed for varying the position of L_2 axially relative to L_1, while a clamping screw enabled the coil L_2 to be fixed when the final adjustment of M was completed. In order to reduce the mutual capacity between the coils the common point D was obtained by joining the adjacent ends of the coils, thus keeping the high-potential ends as far apart as possible.

The most suitable voltage indicator for the purpose is the anode-rectification type of thermionic voltmeter, since, by the use of negative grid bias, grid current can be prevented for the range of voltage it is desired to measure, thus ensuring an infinite resistance between the points A and B. Such a voltmeter can readily be calibrated to measure high-frequency voltages accurately from 0·5 to 1·5 volts. If the resistance to be measured is of the order of 1 or 2 ohms the current I must be appreciable—up to 0·5 ampere—for accurate indication of the voltage, if the added complication of amplifiers is to be avoided. With such currents flowing, the voltage across L_1 will be considerable and, unless M is carefully adjusted, the indicating instrument may be damaged.

The provision of current outputs of the above order does not offer any difficulty with a valve oscillator, but it is important that the entire oscillator, together with the leads to the measuring circuit, be completely screened in order to prevent direct induced voltage into either the secondary of the mutual inductance or the voltmeter circuit. As previously mentioned, this could be checked by reversing the connections of the coupling coil and noting whether this produced an alteration in the apparent resistance of the coil L_1.

The oscillator was enclosed in a copper box having a lid with deep overlaps, while the leads to the measuring circuit were passed through a copper tube soldered into the side of the box. Currents up to 1·0 ampere were obtainable at resonance in the circuit L_1CL_3.

The potentiometer R was of Eureka resistance wire arranged non-inductively by bringing back the return lead close to the potentiometer wire, and was provided with 8 tappings in a total resistance of approximately 2·0 ohms.

The complete circuit connections showing the earthed points are given in Fig. 5. M.S. was a mercury cup

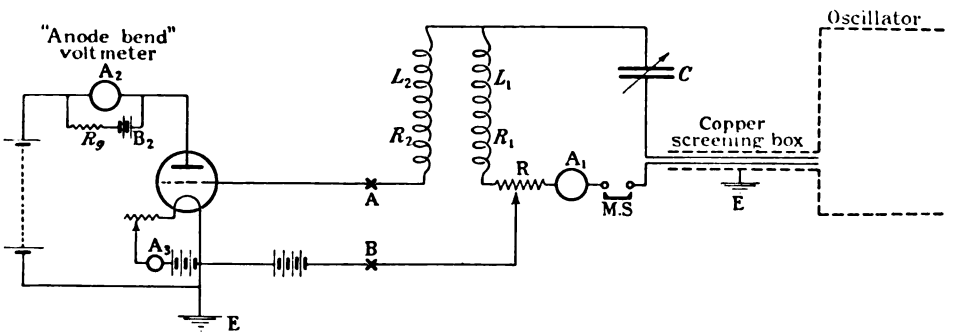

Fig. 5.

switch for opening and closing the circuit L_1CL_3 when required; A_1 a thermo-ammeter for measuring the current in the L_1CL_3 circuit and A_2 a micro-ammeter in the voltmeter plate circuit, the zero deflection being balanced out by a small battery B_2 in series with a high resistance R_g round the micro-ammeter circuit.

Table 1.

	Primary coil	Secondary coil
Turns (closely wound)	25	93
Wire diameter..	0·037 cm	0·031 cm
Wire covering ..	Double cotton	Enamelled
Former diameter	13·3 cm	13·8 cm
D.C. resistance	1·66 ohms	9·03 ohms
Inductance ..	162 μH	1 800μH
Self-capacity ..	By direct measurement $C_1 = 4{\cdot}0\mu\mu$F	Natural wavelength when erected on base of mutual = 210 m $C_2 = 7{\cdot}0\,\mu\mu$F

In order to prevent hand-capacity effects during the measurements, an earthed electrostatic screen, consisting of a number of parallel wires joined together at one end, was erected between the observer and the measuring

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circuit. To enable the mutual inductance to be varied with the screen in position, the control rod of the mutual was made of considerable length and passed through the screen.

The large inductance of the secondary coil was necessary in order to provide the required value of mutual inductance without too close proximity between the primary and secondary coils. When adjusted to

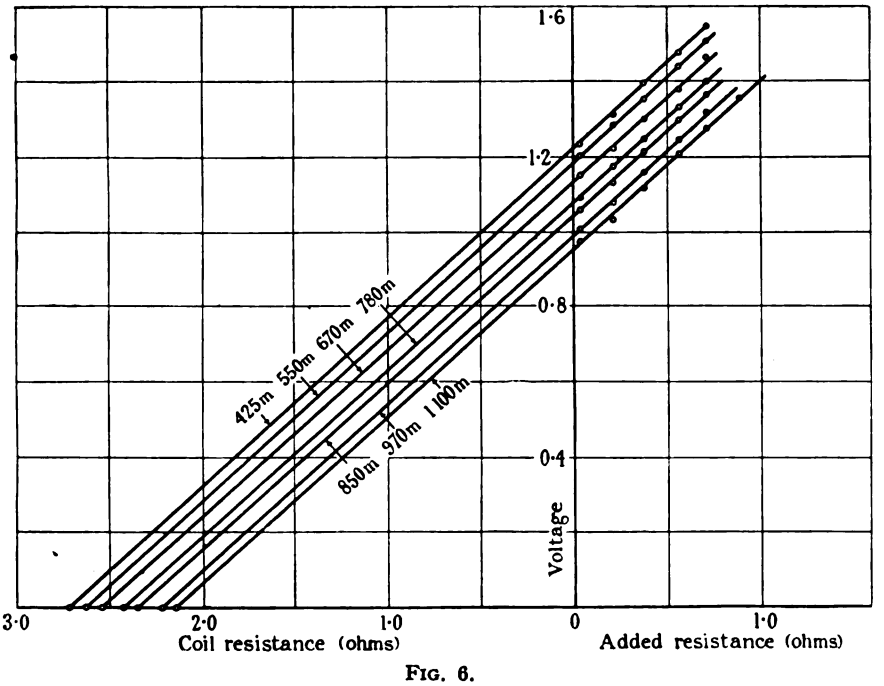

FIG. 6.

(A). The results of a series of measurements, taken over a wavelength range of 1 100 m to 425 m, using this mutual inductance are shown in Fig. 6. The curves connecting voltage between the points A and B, and

give a minimum voltmeter reading the ends of the coils L_1 and L_2 were approximately 1 cm apart. As the frequency was increased it was necessary, in providing the requisite value of M for minimum voltmeter

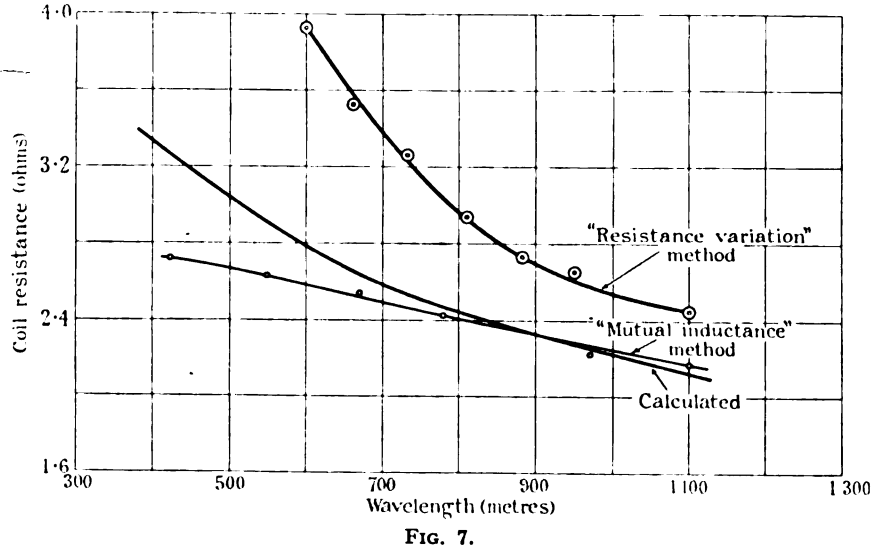

FIG. 7.

added resistance R, when produced backwards to meet the resistance axis, give the apparent resistance R_1 of the primary coil L_1 at the several frequencies.

The constants of the primary and secondary coils of the mutual inductance, as measured independently, are shown in Table 1.

reading, to reduce the coupling between the coils, indicating that capacity effects were present, an effect consistent with equation (6). This became pronounced in the region of 400 m.

Throughout the measurements the current I in the resonant circuit L_1CL_3 was adjusted to 0·43 ampere.

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The apparent resistance R_1 deduced as OV/I (see Fig. 2) agrees to within a few per cent with the derived value OR_1.

A comparison of the above results with the calculated values of resistance using Butterworth's theoretical formula for short single-layer coils is obtained from the curves of Fig. 7. The theoretical formula mentioned is

$$R = R_0\left\{1 + F + uG\frac{d^2}{D^2}\right\}$$

where R_f is the resistance at frequency f cycles per second, R_0 is the resistance to direct current, $(1 + F) = \frac{1}{4}(\sqrt{2z} + 1)$, $G = \frac{1}{8}(\sqrt{2z} - 1)$, $z = \pi d\sqrt{(2f/\rho)}$, d is the wire diameter and D the pitch of the turns, each in centimetres. The factor $u = 3{\cdot}29 + b/a$, where b is the length of the coil and a its radius.

TABLE 2.

	Primary coil	Secondary coil
Turns closely wound	41	136
Wire diameter ..	0·045 cm	0·024 cm
Wire covering ..	Double cotton	Double silk
Paxolin former diameter ..	15·25 cm	15·25 cm
D.C. resistance	2·09 ohms	24·80 ohms
Inductance ..	370 μH	3 830 μH
Self-capacity ..	By direct measurement $C_1 = 4{\cdot}5$ $\mu\mu$F	Natural wavelength when erected on base of mutual = 335 m $C_2 = 8{\cdot}2$ $\mu\mu$F

At the longer wavelengths there is close agreement between the two resistance curves, with the value of resistance as measured by the mutual-inductance method slightly in excess of the calculated values. This is to be expected, since Butterworth's formula is based purely on a consideration of the eddy-current losses in the coil and does not include the effect on the apparent coil resistance of either the capacity between turns or the high-frequency dielectric losses in the coil former. At the shorter wavelengths, however, there is an increasing tendency for the mutual-inductance method to underestimate the coil resistance, indicating that capacity effects, in particular that due to secondary coil self-capacity, are becoming of importance in affecting the derived values of resistance. An estimate of the effect of the self-capacities C_1 and C_2 may be obtained from a consideration of equation (7). By use of the measured coil constants it can be shown that C_1 and C_2 may be expected to produce a net reduction in apparent coil resistance of 1·5 per cent at 1 000 m and 7·5 per cent at 400 m, although this does not take into consideration any compensating effect of mutual capacity, which would become of importance if C_{12} exceeded 2 $\mu\mu$F or 3 $\mu\mu$F.

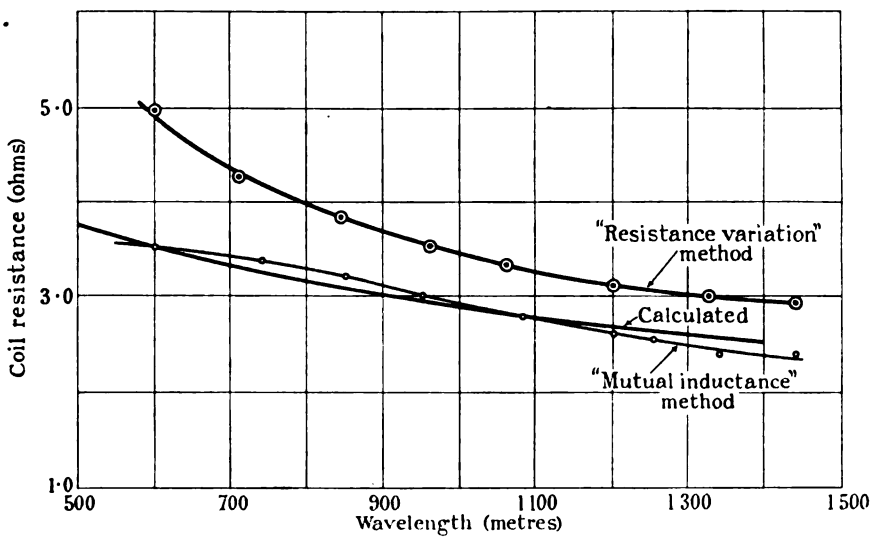

FIG. 8.

The results of high-frequency resistance measurements on the primary coil L_1 when the resistance variation method is employed are also given in Fig. 7. The serious discrepancy between these resistance values and those obtained by the mutual-inductance method is, in part, due to this under-estimating effect. The correction is, however, much less than the difference to be accounted for, and it is therefore justifiable to account for this difference mainly by the unknown resistance of the condenser used in the resonant circuit of the resistance-variation method. This serves to show how important the condenser resistance, if unknown, may be as a source of error in the latter method.

(B). Table 2 gives the constants of a pair of coils wound to replace the primary and secondary of the previous mutual inductance for the purpose of further checking the accuracy of the method. The results of resistance measurements on the primary coil by the mutual inductance and resistance variation methods, together with the calculated values, are shown in Fig. 8.

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The resistance values obtained by the mutual method are again in close agreement with the calculated values, but appreciably low compared with the resistance variation measurements. An estimation of the effect of primary and secondary coil self-capacities on the resistance values obtained by the mutual method shows them to be responsible for a net reduction in apparent

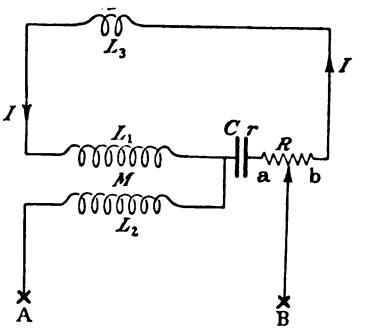

FIG. 9.

resistance of 4·5 per cent at 1 500 m and 15·0 per cent at 700 m. It is worthy of note, however, that a mutual capacity of 6 $\mu\mu$F would be sufficient to compensate entirely for this 15 per cent reduction at 700 m by the introduction of an equal error of reverse sign. While such a high value of C_{12} is not to be expected in the coils of the mutual themselves, since at the setting for minimum voltmeter reading the adjacent ends of the coils L_1 and L_2 were more than 1 cm apart and these ends were joined, it may easily be provided in the input capacity of the voltmeter. This capacity would exist between the points A and B (Fig. 3), and although the valve used in the voltmeter was a V24 low-capacity valve it is unlikely that its capacity would be less than 5·0 $\mu\mu$F.

(6) Conclusions.

The recorded work serves to demonstrate that the method is capable of giving results which compare favourably with the calculated values of coil resistance at wavelengths well removed from the natural wavelength of the secondary coil. At wavelengths approaching this the small self and mutual capacities inherent in the mutual inductance are capable of introducing errors which, while to some extent compensating, may be far from inconsiderable. These errors may, however, be reduced by careful design of the mutual inductance and may be estimated with reasonable accuracy. As an offset to this disadvantage, the method possesses the advantages that no knowledge of the tuning condenser power factor is necessary, and there is no difficulty in manipulation or in indication of the balance condition.

(7) Application to Condenser Resistance Measurement.*

The system shown in Fig. 1 is readily adapted to the measurement of condenser resistance by a modification of the circuit connections to those of Fig. 9.

The condenser C, the resistance r of which is to be measured, is connected in series with the primary coil L_1 and a small coil L_3 coupled to the oscillation generator. The secondary coil L_2 has one end joined to the common point of L_1 and C, and the other to one terminal A of the voltmeter. The other voltmeter terminal B is taken to the potentiometer R connected in series with the L_1CL_3 circuit.

For a current I of frequency $\omega/(2\pi)$, caused to circulate round the circuit L_1CL_3 by tuning to the generator frequency, the voltage between the points A and B with the potentiometer tap at "a" may be written

$$V = I\left\{r + j\left(M\omega - \frac{1}{C\omega}\right)\right\}$$

As M is varied, the voltage V passes through a minimum value V_1 for $M\omega = 1/(C\omega)$, that is for $M = 1/(C\omega^2)$,

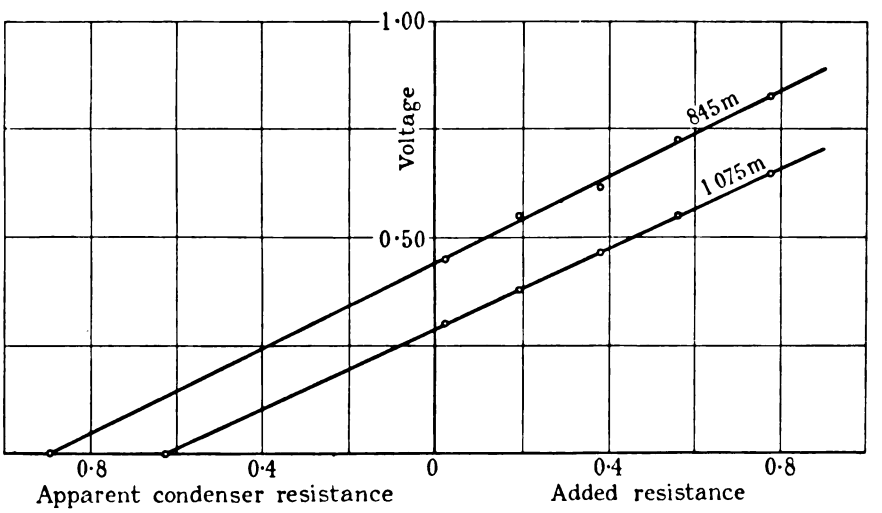

FIG. 10.

* E. B. Moullin, *loc. cit.*, p. 145.

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when $$V_1 = rI \quad . \quad . \quad . \quad . \quad . \quad . \quad (8)$$

For this value of M the insertion of a pure resistance R between the points a and B increases the voltmeter reading to V_2, where

$$V_2 = (r + R)I \quad . \quad . \quad . \quad . \quad . \quad (9)$$

The unknown resistance r may be determined from equations (8) and (9) in terms of V and R, or from a curve connecting V and R obtained by varying R through a series of values.

The curves of Fig. 10 show the results obtained on a large variable air condenser using the mutual inductance, the constants of which are given in Table 1, at wavelengths of 1 075 and 845 m, both well removed from the natural wavelength of the secondary coil. The condenser resistances at these two wavelengths, and for condenser capacities of 1 980 and 1 220 $\mu\mu$F respectively, are thus found to be 0·62 and 0·89 ohm. These values of condenser resistance correspond to condenser power-factors of $2{\cdot}15 \times 10^{-3}$ and $2{\cdot}42 \times 10^{-3}$ respectively.

While it was expected that the power factor would be high, these values seem to be larger than one would justifiably expect in an air condenser.

Avoiding as it does the need for a standard condenser of known power factor, as is required in the comparison methods of condenser-resistance measurement, the method would appear to justify development.

The thanks of the author are due to the Governing Body of the Bradford Technical College for the facilities provided at the College for carrying out the recorded work, and also to Mr. E. B. Moullin, Associate Member, for his interest and for his help in scrutinizing the manuscript.

APPENDIX.

The general expression for the potential difference between the points A and B (Fig. 4), given by equation (4), may be written in the form

$$V = I(a + jb)$$

where

$$a = R_1\left[1 + 2\omega^2\left\{C_1L_1 + C_{12}(L_1 - M)\right\} - \omega^2\left\{C_1M + (R_2/R_1)C_2M - C_{12}\left\{(L_2 - M) + (R_2/R_1)(L_1 - M)\right\}\right\}\right]$$

and

$$b = \omega\left[L_1 + \omega^2\left\{C_1L_1^2 + C_2M^2 + C_{12}(L_1 - M)^2\right\} - (C_1 + C_{12})R_1^2 - M - \omega^2\left\{C_1L_1M + C_2L_2M - C_{12}(L_1 - M)(L_2 - M)\right\} - C_{12}R_1R_2\right]$$

Since in the derivation of equation (1), when self and mutual capacities were not taken account of, it was shown that the voltage V was a minimum when M was equal to L_1, it is justifiable in the above, in which the effects of the small capacities C_1, C_2 and C_{12} are allowed for, to consider $(L_1 - M)$ as small compared with either L_1, L_2, M or $(L_2 - M)$, L_2 being necessarily much greater than L_1, and therefore M, in order to provide the required value of mutual inductance with reasonably loose coupling between L_1 and L_2. Hence as C_1, C_2 and C_{12} are of similar magnitude, $C_{12}(L_1 - M)$ may be neglected in comparison with C_1L_1, since $C_1L_1\omega^2$ is itself a small quantity compared with unity, provided that $\omega/(2\pi)$ does not approach the natural frequency of the primary coil. Also, since R_1 and R_2 need not be of sensibly different magnitude, $(R_2/R_1)(L_1 - M)$ may be neglected in comparison with $(L_2 - M)$. Similarly in the expression for b, $C_{12}(L_1 - M)^2$ can be considered to be negligible in comparison with $(C_1L_1^2 + C_2M^2)$, since under the previous condition $\omega^2(C_1L_1^2 + C_2M^2)$ is itself small compared with L_1. Also, provided that the natural frequency of the secondary coil is not approached, $C_{12}(L_1 - M)(L_2 - M)$ may be neglected in comparison with $(C_1L_1M + C_2L_2M)$.

The expressions for a and b therefore reduce to

$$a = R_1\left\{1 + 2\omega^2C_1L_1 - \omega^2C_1M - \omega^2(R_2/R_1)C_2M + \omega^2C_{12}(L_2 - M)\right\}$$

$$b = \omega\left\{L_1 + \omega^2(C_1L_1^2 + C_2M^2) - (C_1 + C_{12})R_1^2 - M - \omega^2(C_1L_1M + C_2L_2M) - C_{12}R_1R_2\right\}$$

that is

$$a = R_1(1 + 2\omega^2C_1L_1 + \omega^2C_{12}L_2) - M\left[R_1\omega^2\left\{C_1 + (R_2/R_1)C_2 + C_{12}\right\}\right]$$

$$b = \omega\left\{L_1 + \omega^2C_1L_1^2 - (C_1 + C_{12})R_1^2 - C_{12}R_1R_2\right\} - M\left[\omega\left\{1 + \omega^2(C_1L_1 + C_2L_2 - C_2M)\right\}\right]$$

$$= \omega\left\{L_1 + \omega^2C_1L_1^2 - (C_1 + C_{12})R_1^2 - C_{12}R_1R_2\right\} - M\left[\omega\left\{1 + \omega^2\left\{C_1L_1 + C_2(L_2 - L_1) + C_2(L_1 - M)\right\}\right\}\right]$$

In the expression for b, $(C_1 + C_{12})R_1^2$ and $C_{12}R_1R_2$ are negligible in comparison with $\omega^2C_1L_1^2$, R_1 and R_2 being small compared with $L_1\omega$. Further, $C_2(L_1 - M)$ is negligible compared with $C_1L_1 + C_2(L_2 - L_1)$.

Hence $$a = \alpha - M\beta \quad . \quad . \quad . \quad . \quad . \quad . \quad (10)$$

and $$b = \eta - M\gamma \quad . \quad . \quad . \quad . \quad . \quad . \quad (11)$$

where $$\alpha = R_1\left\{1 + 2\omega^2C_1L_1 + \omega^2C_{12}L_2\right\}$$

$$\beta = R_1\omega^2\left\{C_1 + (R_2/R_1)C_2 + C_{12}\right\}$$

$$\eta = \omega L_1(1 + \omega^2C_1L_1)$$

$$\gamma = \omega\left\{1 + \omega^2C_1L_1 + \omega^2C_2(L_2 - L_1)\right\}$$

The voltage $V = I(a + jb)$ will be a minimum with respect to variations in M when

$$\frac{d(a^2 + b^2)}{dM} = 0$$

that is when $$M = \frac{\alpha\beta + \eta\gamma}{\beta^2 + \gamma^2}$$

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On substitution for α, β, η and γ,

$$M = \frac{R_1^2\{C_1 + (R_2/R_1)C_2 + C_{12}\}(1 + 2\omega^2C_1L_1 + \omega^2C_{12}L_2) + L_1(1 + \omega^2C_1L_1)\{1 + \omega^2C_1L_1 + \omega^2C_2(L_2 - L_1)\}}{R_1^2\omega^2\{C_1 + (R_2/R_1)C_2 + C_{12}\}^2 + \{1 + \omega^2C_1L_1 + \omega^2C_2(L_2 - L_1)\}^2}$$

Expanding and neglecting terms of the second order of small quantities—to which

$$R_1^2\{C_1 + (R_2/R_1)C_2 + C_{12}\}(1 + 2\omega^2C_1L_1 + \omega^2C_{12}L_2)$$

and $$R_1^2\omega^2\{C_1 + (R_2/R_1)C_2 + C_{12}\}^2$$

in the numerator and denominator respectively belong —this required value of M reduces to

$$M = \frac{L_1\{1 + 2\omega^2C_1L_1 + \omega^2C_2(L_2 - L_1)\}}{1 + 2\omega^2C_1L_1 + 2\omega^2C_2(L_2 - L_1)}$$
$$= L_1\{1 - \omega^2C_2(L_2 - L_1)\}$$
$$= L_1(1 + L_1C_2\omega^2 - L_2C_2\omega^2)$$

to the first order of small quantities,

The minimum voltage between the points A and B (Fig. 4) is obtained by substituting this value of M in the expression

$$V = I\sqrt{(a^2 + b^2)}$$

Using equation (11) we obtain

$$b = \omega L_1(1 + \omega^2C_1L_1) - \omega L_1\{1 + \omega^2C_1L_1 + \omega^2C_2(L_2 - L_1)\}\{1 - \omega^2C_2(L_2 - L_1)\} = 0$$

to the first order of small quantities.

Hence $V_{min.} = Ia$, where, by equation (10),

$$a = R_1(1 + 2\omega^2C_1L_1 + \omega^2C_{12}L_2) \cdot - R_1\omega^2\{C_1 + (R_2/R_1)C_2 + C_{12}\}L_1$$

to the first order of small quantities

$$= R_1\{1 + \omega^2C_1L_1 - (R_2/R_1)\omega^2C_2L_1 + \omega^2C_{12}(L_2 - L_1)\}$$

Hence, finally,

$$V_{min.} = IR_1\{1 + L_1C_1\omega^2 - (R_2/R_1)L_1C_2\omega^2 + (L_2 - L_1)C_{12}\omega^2\}$$

MEETINGS OF THE WIRELESS SECTION.

78TH MEETING OF THE WIRELESS SECTION, 6TH NOVEMBER, 1929.

Colonel Sir T. F. Purves, O.B.E., President, took the chair at 6 p.m.

The minutes of the meeting of the Wireless Section held on the 1st May, 1929, were taken as read and were confirmed and signed.

A vote of thanks to Commander J. A. Slee for his services as Chairman of the Wireless Section during the session 1928–1929, proposed by Lieut.-Col. A. G. Lee, O.B.E., M.C., B.Sc., and seconded by Major B. Binyon, O.B.E., M.A., was carried with acclamation.

Captain C. E. Kennedy-Purvis, R.N., Chairman of the Wireless Section, then delivered his Address (see page 1).

A vote of thanks to Captain Kennedy-Purvis for his Address was proposed by the President and, after being put to the meeting, was carried with acclamation.

The meeting terminated at 7 p.m.

79TH MEETING OF THE WIRELESS SECTION, 4TH DECEMBER, 1929.

Captain C. E. Kennedy-Purvis, R.N., Chairman of the Section, took the chair at 6 p.m.

The minutes of the meeting of the Wireless Section held on the 6th November, 1929, were taken as read and were confirmed and signed.

A paper by Mr. G. Shearing, B.Sc., and Captain J. W. S. Dorling, R.N., Members, entitled "Naval Wireless Telegraph Communications" (see page 9), was read and discussed.

The meeting terminated at 7.55 p.m. with a vote of thanks to the authors, which was moved by the Chairman and carried with acclamation.

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A METHOD OF MEASURING THE OVERALL PERFORMANCE OF RADIO RECEIVERS.*

By H. A. THOMAS, M.Sc., Associate Member.

[From the National Physical Laboratory: communicated by permission of the Radio Research Board.]

(*Paper first received 20th June, 1929, and in final form 13th March, 1930; read before the* WIRELESS SECTION *15th January, 1930.*)

SUMMARY.

The paper describes researches which have been carried out at the National Physical Laboratory for the Radio Research Board. An improved method of measuring the amplification of a radio receiver is described, together with the manner in which this method can be applied to practical cases, particularly at short wavelengths. The apparatus covers a wavelength range of from 15 to 30 000 metres, and is capable of dealing with inputs of from 0·25 to 1 000 mV over this range. After describing in detail the construction and operation of the apparatus in Section (1), the results of tests on four receivers of widely different types are given in Section (2), and some general deductions are obtained from the properties of these receivers. In conclusion, suggestions are made in Section (3) for a possible means of classifying receivers for different purposes.

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Section (1).—Method of measurement.
- (*a*) Introduction.
- (*b*) The construction of the screened cabin for the oscillator.
- (*c*) The construction of the screened input resistance.
- (*d*) The input oscillator.
- (*e*) The measurement of the output voltage.
- (*f*) The complete circuit arrangements.

Section (2).—Examples of measurements upon various types of receivers.
- (*a*) Introduction.
- (*b*) A 4-valve receiver (200–16 000 m).
- (*c*) A commercial short-wave receiver (15–60 m).
- (*d*) A screened short-wave receiver (15–80 m).
- (*e*) A supersonic-heterodyne receiver with frame aerial (50–500 m).

Section (3).—Suggestions for specifying the performance of receivers.

SECTION (1).—METHOD OF MEASUREMENT.

(*a*) *Introduction.*—The aim of all measurements of the constituent parts of a radio receiver is ultimately to obtain a characteristic relationship between the input supplied to the receiver and the output voltage given by the receiver, together with an analysis of the behaviour of the individual portions of the apparatus. Such a measurement is the most important of all those that can be obtained, and at the same time is the most difficult. The difficulty lies in the fact that the output quantities are in general very large compared with the input quantities, so that the slightest form of accidental coupling between the input and the output circuits introduced by the measuring apparatus may completely modify the behaviour of the receiver. Another difficulty lies in the fact that the input and output frequencies are dissimilar.

Most of the early attempts to obtain the overall characteristics of a receiver were based on an assumption that the overall performance could be represented in terms of the product of the performances of the separate components. If this assumption were valid, the problem would be simple, but, unfortunately, the effects due to the input impedance of an amplifying stage, together with the various forms of retroaction, render this assumption unjustifiable in all but the very simplest cases.

Due to the difference in frequency between the input and output, the large magnitude of the step-up ratio, and the retroaction effects, substitution, bridge and comparison methods are difficult to employ. The input must be supplied by means of an artificially generated signal, the form of which must be such that the results obtained can be readily applied to the actual conditions met with in practice.

The relatively large output must also be measured by a method which in no way disturbs the normal operation of the receiver.

An overall performance measurement implies the inclusion of the input tuned circuit, the antenna or frame-coil, whichever is used in practice with the receiver. The problem of applying a known e.m.f. to this tuned circuit may be solved in one of the following three ways:—

(i) By inducing the e.m.f. into the main tuning inductance by means of a known mutual inductance, the current in the primary of which is known.

(ii) By passing a known current through a condenser of known value which forms part of the tuning capacity.

(iii) By inserting in series with the tuned circuit a small non-inductive resistance or a small inductance carrying a known current.

The first method has been used,† but it involves the measurement of a mutual inductance of large ratio at radio frequencies. It is difficult to make such measurements with any accuracy, and also the tuning circuit of the receiver must always include, or consist of, this secondary circuit of the calibrated mutual inductance, which must always be carefully screened from the rest

* Reprinted from *Journal I.E.E.*, 1930, vol 68, p. 475.

† T. A. SMITH and G. RODWIN: "Notes on Radio Receiver Measurement," *Proceedings of the Institute of Radio Engineers*, 1927, vol. 15, p. 387.

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of the receiver. The method is therefore inapplicable as a general method of testing receivers in which the tuned circuit forms part of the complete arrangement.

In the second method, since the value of the tuning condenser is specified by the wavelength range, the e.m.f. which can be applied can only be controlled by the current passed through this condenser and supplied by the main oscillator. Consequently the possible range of the input e.m.f. is very limited. For the smallest measurable radio-frequency current, the value of this condenser must be very large if the e.m.f. across it is to be of the right order.

is connected to the joint between the resistance wire and the concentric return, the batteries will be connected to the amplifier screen which is at earth potential, which condition is required for the imitation of the normal operation of the receiver. It is important that this battery earthing connection be adhered to, otherwise the capacities to the screen may modify the measured performance. If the amplification is very large, the resistance wire becomes very short and at very high frequencies its inductance becomes important. This inductance is too small to measure and must be calculated from the known dimensions of the wire.

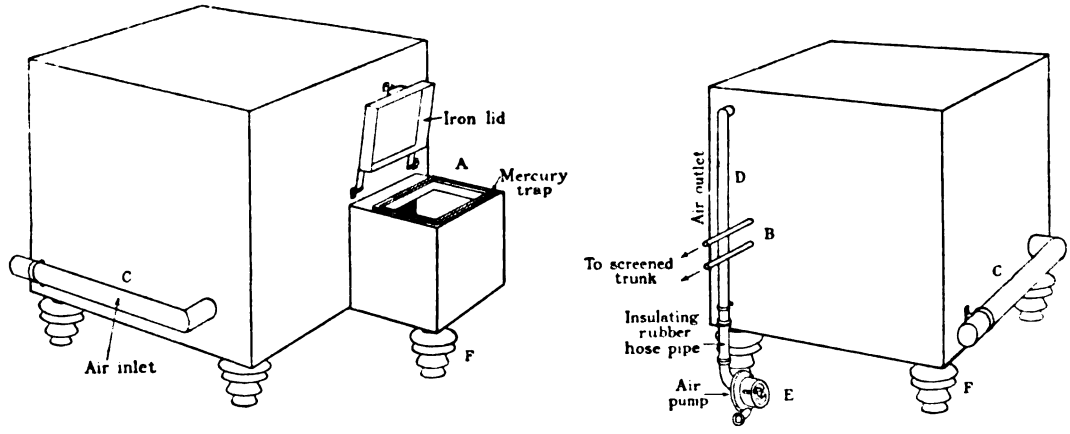

FIG. 1.—Screened oscillator cabin.

The third method has been adopted to a greater extent than the previous methods,* owing to its more general application and its comparative independence of frequency. Small non-inductive resistances can be made within fairly wide limits and, with a reasonable range of current, can be used for a wide range of input voltage. A radio-frequency oscillation, modulated to a known extent at an audio frequency, supplies current to these resistances. The prevention of direct magnetic and electric induction from the oscillators and leads to the resistance demands the most elaborate screening of these components. This is effected by enclosing the oscillators in a completely screened metal room and screening the leads to the resistance. The amplifier, too, must be housed in a screened room to prevent induction from external sources. Refinement in measurement is dependent upon the efficacy of these screens, and practically nothing except complete shielding is permissible.

The screened resistance takes the form of a straight resistance wire with a concentric return, by which means complete shielding is effected. If the resistance wire is tapped at various points, e.m.f.'s of known magnitudes are obtained. At very high radio frequencies it is important that the current through the wire should be measured at a point as near to the tapping as possible. If the filament of the amplifier

In practice, the only factors limiting the accuracy of the measurement are:—

(i) The impossibility of measuring the effective impedance of the screened input resistance, and the consequent dependence of this value upon the calculated value of the inductance of the wire. The possibilities of error, however, are small.

(ii) The difficulty of measuring the current in the resistance at very high radio frequencies. At present a non-contact thermo-junction ammeter is used, and the assumption is made that its d.c. calibration is valid at all frequencies.

This limitation is, however, imposed on many other measurements at high radio frequencies at the present time. A better method of measuring very high radio-frequency currents would naturally improve the accuracy of the measurements at such frequencies.

(*b*) *The Construction of the Screened Cabin for the Oscillator.*—A perspective view of the completed oscillator cabin is shown in Fig. 1. The screened cabin is a $6\frac{1}{2}$ ft. cube and consists of a wooden skeleton covered with No. 22 S.W.G. tinned iron sheet. Great care was taken to solder over every joint, and the completed box was hermetically airtight. Previous experience with screened doors had shown the difficulty of making them sufficiently effective, and consequently a mercury-sealed trap door (shown at A) was adopted.* The seal consists of a welded angle-iron frame dipping into a

* H. A. THOMAS: "The Performance of Amplifiers," *Journal I.E.E.*, 1926, vol. 64, p. 253, and *Proceedings of the Wireless Section*, 1926, vol. 1, p. 12; E. KLOTZ: "The Quantitative Measurement of Receivers," *Telefunkenzeitung*, 1927, vol. 8, p. 54; H. D. OAKLEY: "Quantitative Determination of Radio Receiver Performance," *Journal of the American I.E.E.*, 1927, vol. 46, p. 498.

* R. L. SMITH-ROSE: "On the Electromagnetic Screening of a Triode Oscillator," *Proceedings of the Physical Society of London*, 1922, vol. 34, p. 127.

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channel-shaped iron frame containing mercury. This door has been tested and found to be entirely satisfactory, but it has been found important to maintain the portion of the iron frame immersed in the mercury in a thoroughly clean condition. The output current leads are taken through seamless steel tubing (shown at B).

The ventilation of such a box presents some difficulty, the method here adopted making use of two long copper pipes C and D, provided with three copper-gauze baffles which give adequate screening while permitting a sufficient passage for air. Pipe D is connected by an insulating rubber hose-pipe to a centrifugal pump E, and pipe C is the inlet. The diameter of pipe D is much smaller than that of C, so that the air pressure inside

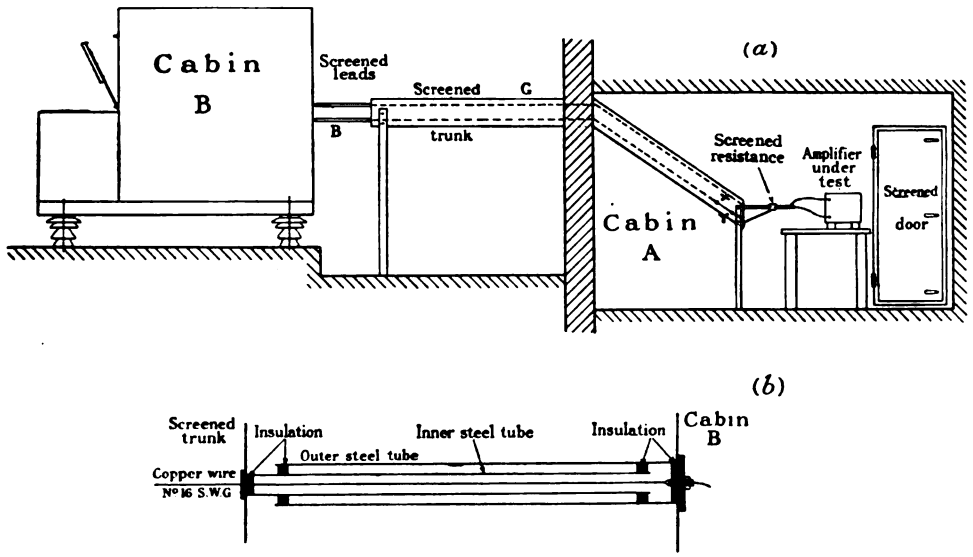

FIG. 2.

the box is only slightly lower than that of the outer atmosphere. The complete box is insulated from earth by means of the five large insulators F.

The output pipes B are insulated from the connecting trunk G, Fig. 2 (*a*), by a method, shown in detail in Fig. 2 (*b*), which gives good insulation and at the same time effective screening. Tests made with the completed box have shown that it is sufficiently screened for the purposes required.

(*c*) *The Construction of the Screened Input Resistance.*—Since the radio-frequency current is taken through the trunks to cabin A containing the amplifier, the radio-frequency resistance needs to be very efficiently screened.

For short-wave work, usually below 40 m, the inductance of the resistance wire is not negligible and consequently it was decided to use a screened straight wire in which the inductance was calculable. Such an arrangement would take the form shown in Fig. 3 (*a*), the current passing through the central wire and back through the screening cylindrical tube. The external field would be negligible, and if a certain length of wire l_1 were used the potential difference across this length could be applied to the amplifier.

The input lead B would have a length l_2 equal to the radius of the cylinder in the magnetic field due to the inner wire, and an e.m.f. could be induced in this wire The construction Fig. 3 (*b*) is better, since the length of unscreened lead B is now materially reduced. However, the distance d must be reduced to a minimum if the capacity between the wire and the screen is to be negligible, such capacity acting as a shunt to the resistance, but this distance d also affects the capacity of the tapping lead to the screen; a compromise is therefore necessary which gives a new minimum capacity shunt effect. Since one lead A of the potentiometer is connected to the screen, it follows that this lead is effectively earthed and is connected to the main screen of cabin A. In all tests this lead is connected to the filament of the amplifier, thus "earthing" all batteries, which condition is normal. Therefore lead Y, Fig. 2, is connected to the screen and need not really be used, but in certain special cases, such as the input to a push-pull stage, no earth is required, and then the lead Y can be disconnected from the screen by the switch and the lead brought out of the screen.

The final arrangement adopted for the input resistance is shown in Fig. 3 (*c*). No. 47 S.W.G. eureka wire is used as the resistance, and No. 40 S.W.G. copper wire is used to connect to the wire X, Fig. 2. The details of the insulation and screening are clearly shown in the diagram. Six such resistances have been made, having values of 0, 0·76, 0·96, 5·139, 11·23 and 27·86 ohms respectively. The resistance was measured by a potentiometer, this obviously being more satisfactory than a bridge resistance measurement, owing to the unknown electrical nature of the soldered joints. The inductance of the six wires was calculated and was found to be $0, 1·39 \times 10^{-3}, 3·39 \times 10^{-3}, 25·98 \times 10^{-3}, 60·45 \times 10^{-3}$ and $176·5 \times 10^{-3}$ μH respectively, these figures being obtained from the lengths of the eureka resistance wire measured by a travelling microscope.

The screened resistance can be plugged into a bayonet socket adjacent to the screened vacuum

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junction, as shown in Fig. 3 (*d*). This junction is of the non-contact type and measures the current within a few inches of where it passes through the resistance wire; this point is of importance since, at very high frequencies, the current varies along the trunk leads.

resistance, but lead B is connected to the screen end of the wire. This resistance is used to ascertain whether any electric or magnetic induction effects are present. Using this resistance, no output should be given if there are no induction effects.

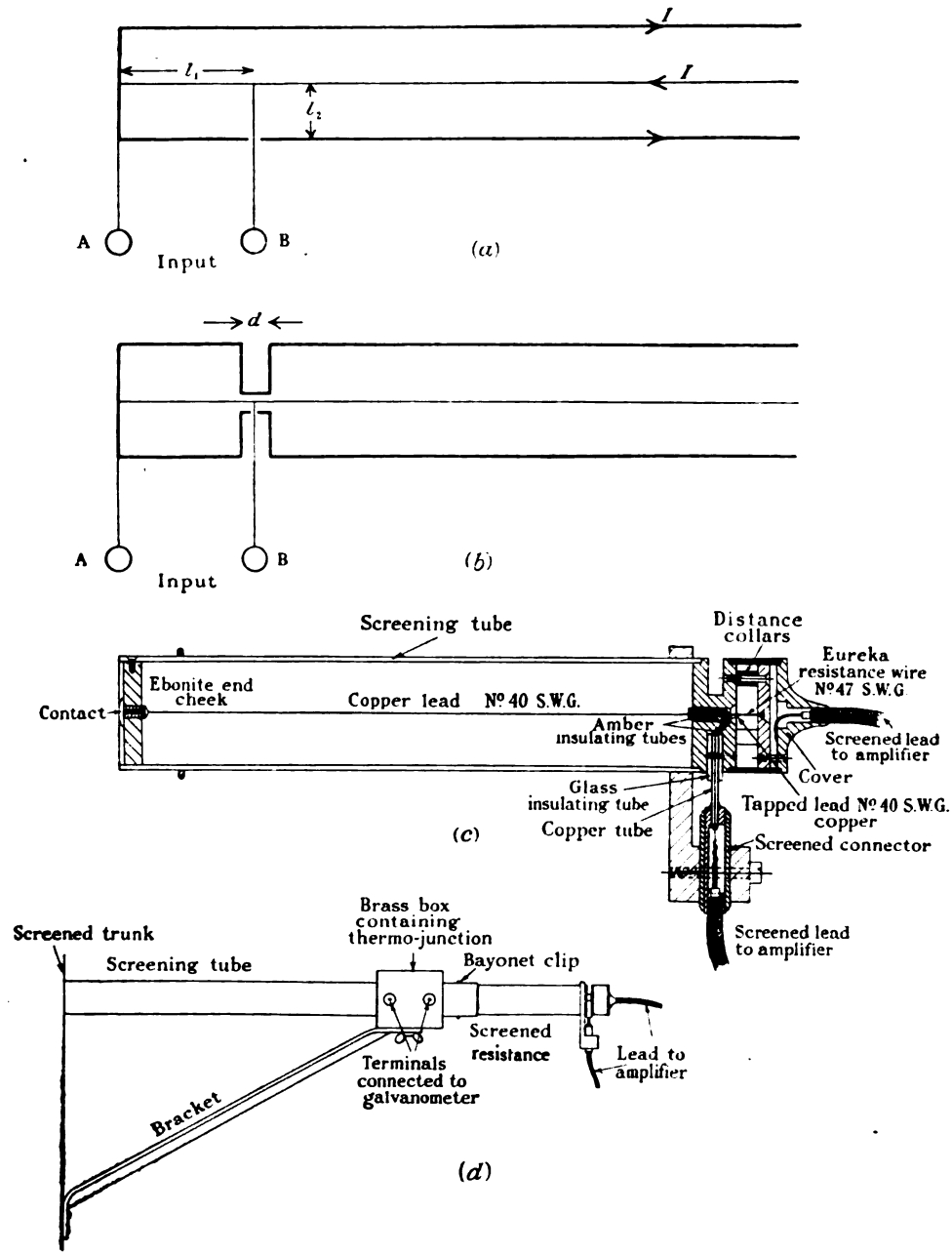

FIG. 3.—Details of screened resistance for supplying known potential differences to a wireless receiver.

The two input leads are screened up to the terminals of the amplifier, the capacity of these leads being insufficient to alter appreciably the impedance of the resistance even at frequencies corresponding to wavelengths in the neighbourhood of 30 m (10 megacycles per sec.). The zero-resistance wire has a construction identical with that of the 0·76-ohm screened

The inductive reactance of the resistances is negligible at frequencies below 10 megacycles, and at frequencies above this value the reactance can easily be calculated. These resistances appear to be quite satisfactory and have been used on wavelengths of 15 m (20 megacycles per sec.).

(*d*) *The Input Oscillator.*—Various types of radio-

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frequency oscillator have been tried, the push-pull type being used for the preliminary work. It has been found, however, that the series-fed Hartley oscillator with grid condenser and leak gives the best results, as this oscillator will work satisfactorily with a 2-volt low-impedance valve and a high-tension supply of 20 volts. Using this circuit arrangement, an oscillator has been built which will cover all wavelengths down to 5 m (60 megacycles per sec.).

In the early work, capacitative coupling to the input screened leads and resistance was adopted, a variable condenser connected between the free potential lead and one end of the oscillating coil being used, while the earthed lead was connected to the centre tapping. The adjustment of this condenser provided a very simple means of varying the current, but unfortunately this type of coupling exaggerates the harmonics and, since the thermo-junction measures the R.M.S. value of the current in the resistance, has been abandoned. The present arrangement employs a tuned circuit inductively coupled to the oscillator, the screened resistance being in series with this circuit. Although not so simple to operate, the purity of wave-form obtained by this method compensates for the increased practical difficulties.

The audio-frequency oscillator consists of a push-pull arrangement with a tuned output circuit in series with the anode feed to the radio-frequency oscillator. The wave-form obtained with this arrangement is sufficiently good for most measurement purposes. The modulation percentage was originally measured by means of an electrostatic voltmeter, but this instrument has now been replaced by a valve voltmeter owing to the sluggishness of the electrostatic instrument. The complete oscillator unit is carefully screened and consists of the audio-frequency oscillator, a modulator unit, the modulation measuring voltmeter, the radio-frequency oscillator and the coupling unit. The complete circuit arrangements are dealt with later in sub-section (*f*).

(*e*) *The Measurement of the Output Voltage.*—Most types of receiver give an output at an audio frequency, and in the early work a vibration galvanometer was used to measure this output. Such an instrument is eminently suitable for the measurement of very small outputs such as those given in telephones but, with more modern receivers, outputs of loud-speaker intensity are more common and the potential on the grid of the power stage is of the order of several volts.

Since the power stage must always be looked upon as a converting mechanism, the amplification must be expressed in terms of the voltage output delivered to the grid of the power stage with respect to some input voltage. To measure reasonably large outputs it is usual to use a valve voltmeter connected across the grid and filament of this stage, but difficulties have been encountered with the use of a valve voltmeter connected in this manner, owing to the large shunt load imposed by this instrument. This effect is illustrated by the results given in Fig. 4, the output as measured by a valve voltmeter being plotted against the input supplied to a 2-stage resistance-capacity amplifier. In curve A, the valve voltmeter was connected directly across the input of the power valve and it is seen that, as the input is increased, the output appears to reach a saturation value due to the load of the voltmeter. If, however, the voltmeter is connected across a 100-ohm resistance inserted in the anode circuit of the power valve, the characteristic is given by curve B, which shows a linear relationship between the input and the output. This low anode-circuit resistance is negligible in comparison with the loud-speaker impedance and consequently its effect, considered as a modifying element in the circuit arrangement, is negligible. The amplification will appear to be higher than that obtained when the voltmeter is connected directly across the input of the power stage when very small outputs are being obtained, as shown by the tangent drawn to the lower part of curve A. Since the modulation percentage

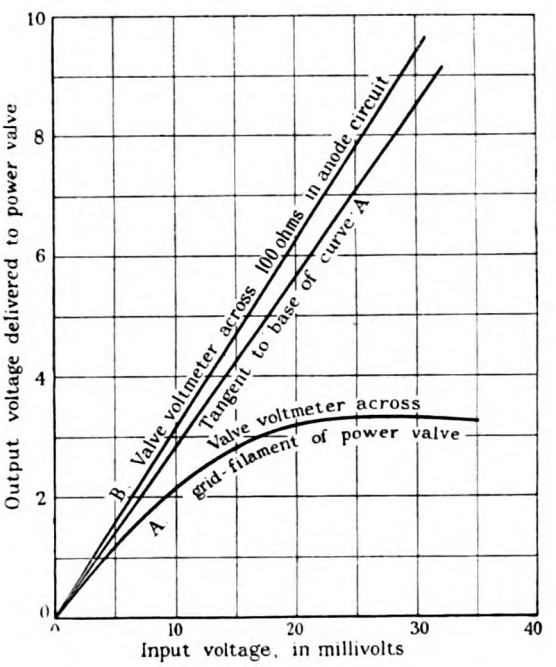

FIG. 4.

never exceeded 15, the only important harmonic produced due to rectification is the second, and this can never produce an error greater than 3 per cent in the output measurement.* In all measurements, therefore, the voltmeter was connected across a low non-inductive resistance in the anode circuit, and the power stage was calibrated by supplying an audio-frequency oscillation directly to the grid and filament by means of a high-resistance potentiometer. By this means the actual voltage on the grid of the power valve can be obtained by noting the deflection of the valve voltmeter connected in the anode circuit.

(*f*) *The Complete Circuit Arrangements.*—Fig. 5 shows the completed arrangement of the apparatus. The radio-frequency oscillation is modulated by means of the screened audio-frequency oscillator and the modulator. By means of a variable coupling between the audio-frequency tuned coupling circuit and the oscillator,

* W. B. MEDLAM and V. A. OSCHWALD: "The Performance of Valve Detectors," *Journal of the Institute of Wireless Technology*, 1928, vol. 1, p. 147.

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and also by shunts across the output transformer of the modulator valve, the modulation percentage may be varied.

It has been found that the output current given by the radio-frequency oscillator is sensibly proportional to $(V - K)$, where V is the high-tension d.c. voltage and K is an experimental constant. If the R.M.S. voltage of the modulator output is E, then the modulation percentage is given by

$$\frac{\sqrt{2}\,E}{V - K} \times 100 \text{ per cent}$$

An inductively-coupled tuned system supplies the screened resistance, the current in which is measured by the non-contact thermo-junction Q. A second junction P is inserted inside the oscillator cabin so that the operator can tune the circuits, but no measurements of current are taken on this instrument. With the switches in cabin A at position (1), the input is normally applied to the amplifier and the negative filament leg is connected to the screens and may be earthed if required; at position (2) the power stage is calibrated with an audio-frequency input by means of a high resistance. The milliammeter in the anode circuit of the power stage is used as an indicator to obtain the same grid bias under both conditions, the bias voltage being controlled by a potentiometer as shown in the diagram.

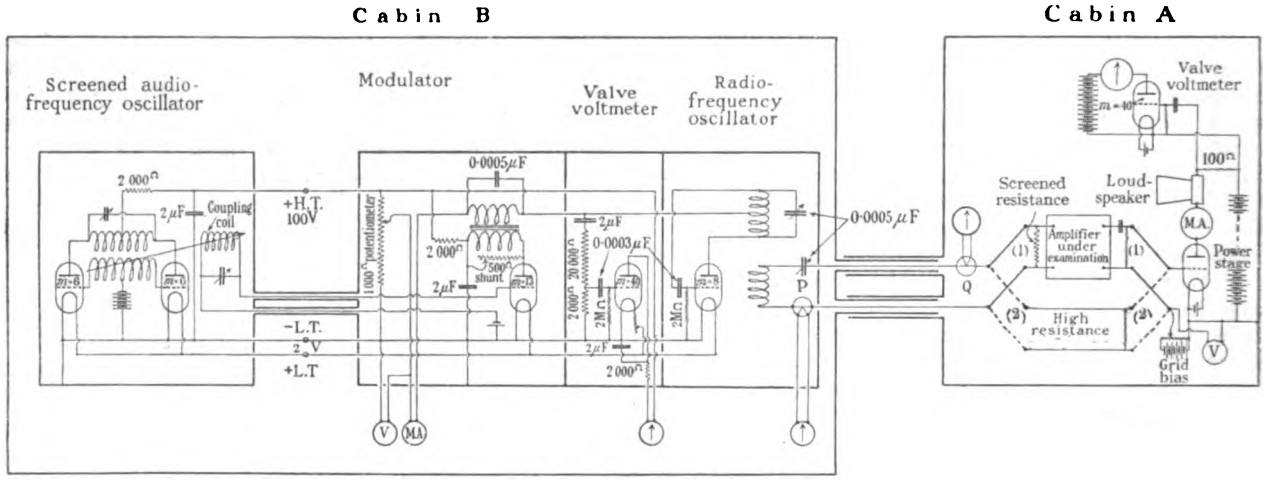

FIG. 5.—Circuit diagram of apparatus for testing the performance of radio receivers

The apparatus is capable of giving an input range of 0·25–1 000 mV at any wavelength from 5–30 000 m (20–0·01 megacycles per sec.), and an output range of 20–6 000 mV at any audio frequency.

SECTION (2).—EXAMPLES OF MEASUREMENTS UPON VARIOUS TYPES OF RECEIVERS.

(*a*) *Introduction.*—Having described the principal features of the measuring apparatus, some results of tests performed upon various types of receivers will now be given. The object of these tests was essentially to demonstrate the utility and range of the apparatus, and, although the results show the properties of the different receivers tested, no mathematical analysis of

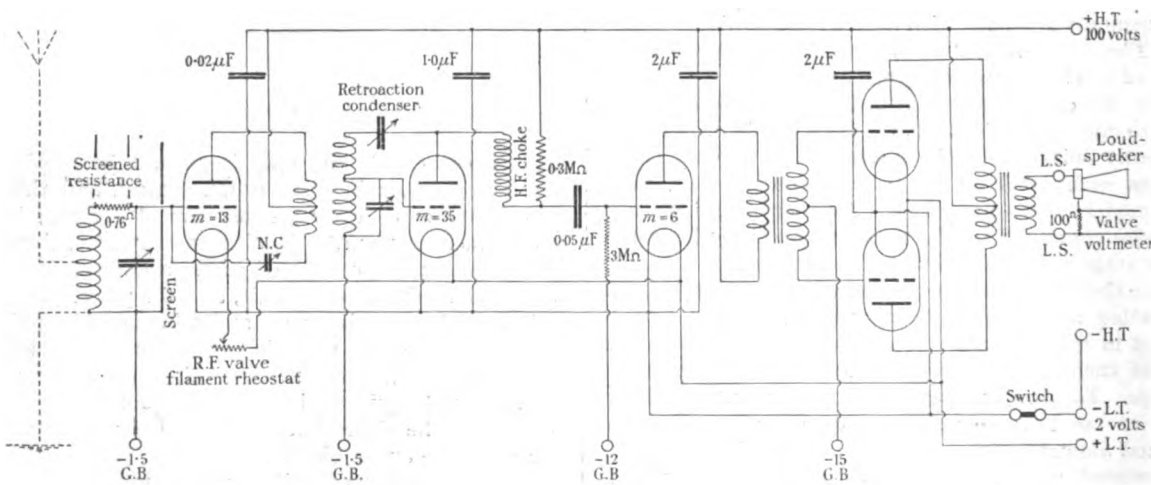

FIG. 6.—Circuit diagram of receiver, showing method of measurement.

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them has been attempted. For the purpose of analytical work, special amplifiers would need to be built and their constants would require exact measurement. The results, however, demonstrate the types of overall

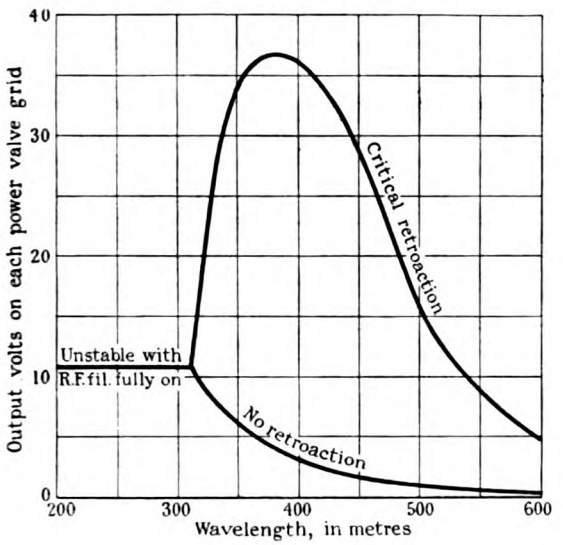

FIG. 7.

characteristic likely to be obtained from receivers of a similar type to those employed.

The receivers used for the tests were deliberately selected to cover several usual types, and, since the

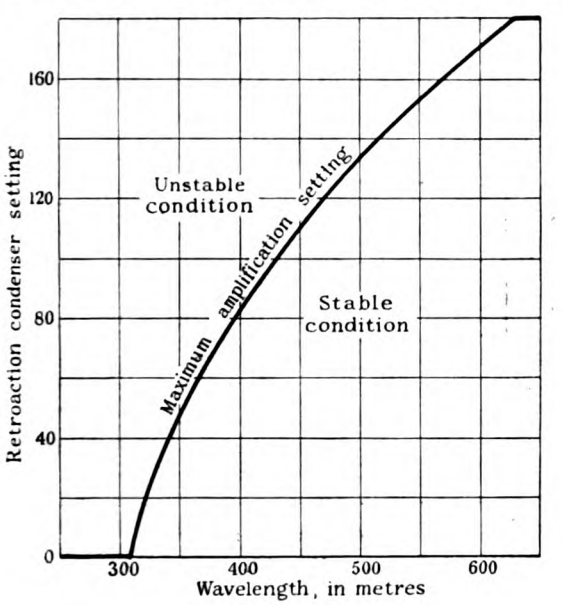

FIG. 8.

apparatus has been proved satisfactory for measuring the performance of these essentially different types, such measurements constitute a reasonable ground for believing that the apparatus can be employed to examine practically any type of receiver.

(*b*) *A 4-valve Receiver* (200–16 000 *m*).—A typical modern 4-valve receiver was examined. The particular receiver chosen had one radio-frequency stage, an anode-current rectifier with resistance-capacity coupling to the first audio-frequency stage, a transformer-coupled audio-frequency stage, and finally a push-pull output stage. The circuit arrangement is shown in Fig. 6. The radio-frequency stage is neutrodyned and a tapped secondary winding on the radio-frequency transformer provides the retroaction. The rectifier has a choke in series with the anode resistance, and the normal aerial and earth connections are shown by the dotted lines. For the purposes of the test the normal earth connection was taken to the screen, which was effectively earthed, and the known input voltage was applied to the tuned circuit as shown.

The valve voltmeter was connected across a 100-ohm resistance in series with the loud-speaker. To calibrate the push-pull output valves, the tapped secondary was

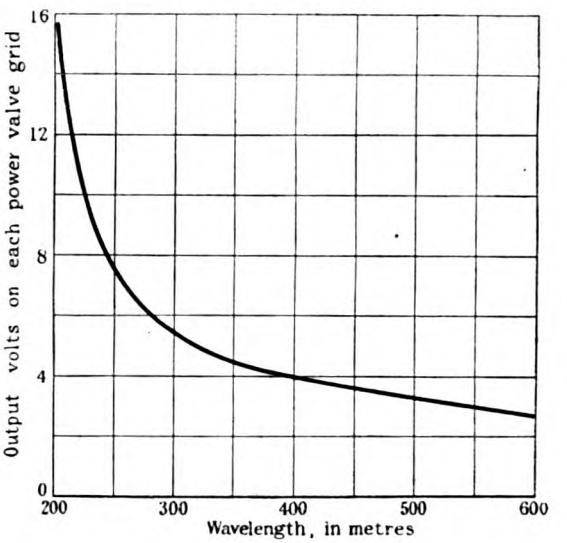

FIG. 9.

removed and a tapped non-inductive resistance was substituted, the middle point being earthed. The audio-frequency oscillator fed this resistance and gave the output on the valve voltmeter.

An earthed screen is used to prevent retroaction between the two radio-frequency circuits. The current in the filament of the radio-frequency valve is adjustable by means of a rheostat, this being used as a volume control and also as a stabilizer at certain wavelengths where the neutrodyne arrangement is not satisfactory.

A constant input of 5 mV modulated 15 per cent at 1 000 cycles per sec. was applied to the tuned circuit at various frequencies, and the R.M.S. output voltage delivered to either grid of the push-pull output valves was measured with the filament current of the radio-frequency valve at maximum and with the retroaction condenser at zero. This characteristic is shown in Fig. 7. At a wavelength of 310 m the receiver became unstable, and at any wavelength below this the radio-frequency valve filament-current had to be reduced to maintain stability. The amplification at these lower

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wavelengths could not be raised above the value given at 310 m. At wavelengths above 310 m the retroaction condenser was now adjusted to the threshold of oscillation and the output was measured. It will be seen from Fig. 7 that the maximum amplification occurred at a wavelength of 380 m. Any point lying between the two curves represents a possible stable condition depending upon the setting of the retroaction condenser. The actual settings of this condenser are shown in Fig. 8. Below 310 m the receiver is unstable with maximum filament current in the radio-frequency valve, even with the retroaction condenser at zero, whereas above 625 m the receiver is stable when all the extra retroaction is inserted. All points lying below the curve represent a condition of stability, and all points above the curve one of instability.

The next test was made at the constant wavelength of 370 m with a constant modulation of 10 per cent but

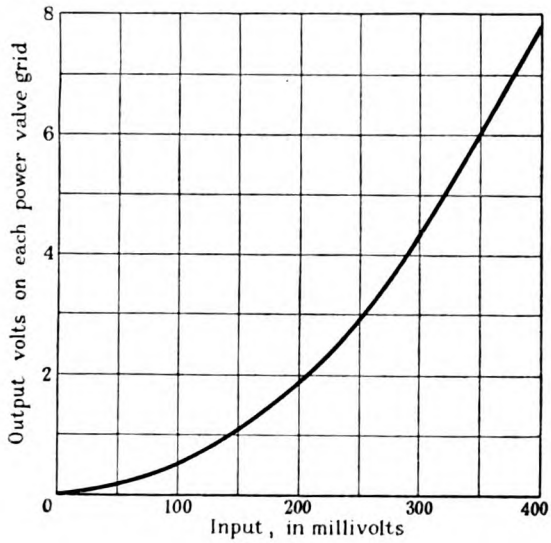

FIG. 10.

with variable input. The relationship between the output and the input was found to be very nearly that of a square law. This test was made at different wavelengths and always gave a similar relationship, from which we deduce that this characteristic represents the property of the rectifier, since the radio-frequency and audio-frequency portions of the receiver sensibly obey a linear law.

It was found that if the output exceeds about 4 volts on each power-valve grid, the relationship between the output and the modulation percentage departs from a linear law. This has been found to be generally true, irrespective of wavelength, and is clearly due to some saturation effect.

A constant input of 70 mV was now applied directly to the radio-frequency stage and the circuit was tuned to give the maximum amplification with no extra retroaction and with the full filament current at each wavelength. The output is plotted in Fig. 9 and it is seen that the radio-frequency stage is stable down to a wavelength of 200 m. The rectifier characteristic obtained by applying the known e.m.f. to the rectifier grid is shown in Fig. 10, the input being applied directly to the rectifier. The relationship between the input and output voltages for a definite modulation percentage is a perfect square law. From these two characteristics, Figs. 9 and 10, the amplification of the radio-frequency

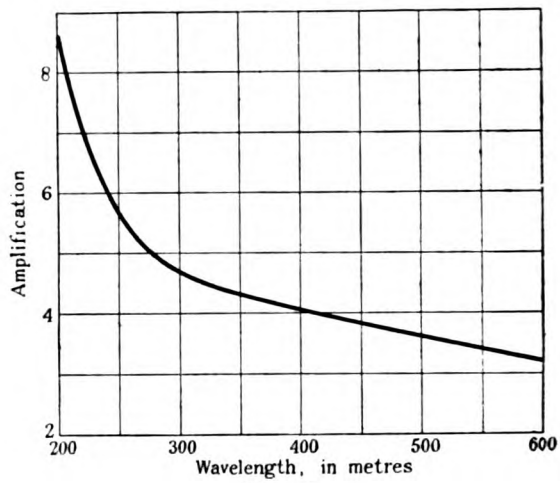

FIG. 11.

stage can be obtained and is plotted in Fig. 11. An examination of this figure shows that the amplification of the radio-frequency stage rises from about 3·2 at 600 m to nearly 9 at 200 m. The receiver is unstable at this wavelength if an input circuit is connected before

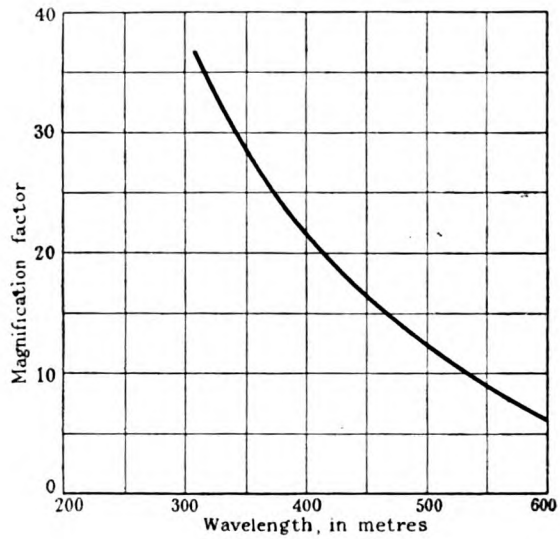

FIG. 12.

the radio-frequency stage, but the figure represents the stable condition with no input circuit. Since the voltage factor of the radio-frequency valve is 13, this amplification figure is very low. Tests on other radio-frequency amplifiers have given similar low values, and, where high values are obtained approaching the valve voltage-factor, retroaction is usually responsible for such values.

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By combining the overall performance characteristic obtained by supplying the tuned input circuit with a known signal, and the radio-frequency valve characteristic, we can obtain a figure for the magnification factor or coil factor of the tuned input circuit; this is shown in Fig. 12. In the absence of all loading and retroaction effects, this characteristic should be given by $\omega L/R$, where L and R are respectively the inductance and radio-frequency resistance of the input tuned circuit.

An attempt was made to reconcile the experimental figures for the amplification of the radio-frequency stage and the magnification factor of the input circuit with the theoretical figures obtained by a cursory examination of the circuit. The radio-frequency resistance of the input circuit and the radio-frequency stage

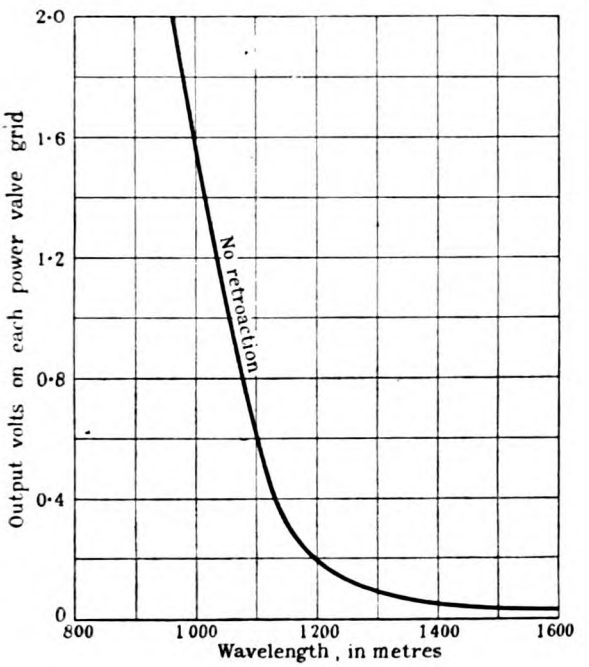

FIG. 13.

load circuit was obtained by measurement, together with the inductance and self-capacity of the coils used in these circuits. The simplest examination of these figures revealed the serious discrepancy between the experimental results and the results obtained by any simple theoretical calculation which neglects the effects due to retroaction and input impedance: not only are both experimental figures far lower than the predicted figures obtained by such a simple analysis, but the form of the curves is different. The departure in form shows that a proportionately higher figure is obtained as the frequency is raised, which is undoubtedly due to retroaction effects at the higher frequencies. It is clear that the input impedance and retroaction effects are very serious and no useful results would be obtained by a closer examination of the characteristics, as the receiver is one in which no special precautions have been taken to prevent stray retroaction. The results given are of considerable interest, in that they demonstrate the general properties obtained with a typical receiver, but a closer examination would be impossible on account of the numerous unknown quantities involved.

The properties of the receiver on long waves, using two different coils, are shown in Figs. 13, 14 and 15. In Fig. 13 the overall performance characteristic is

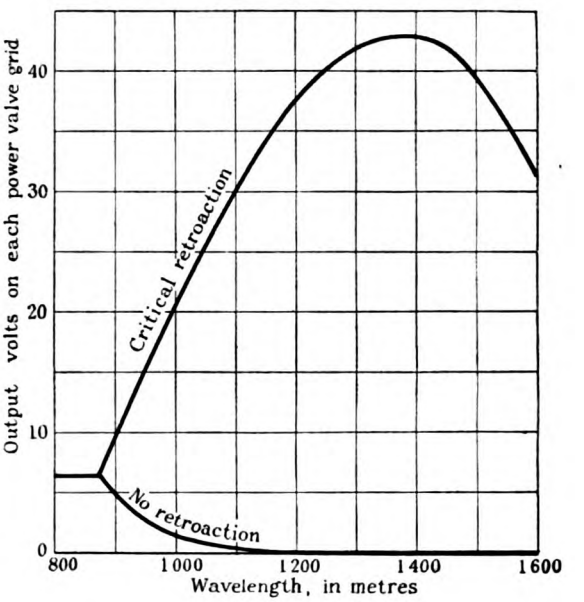

FIG. 14.

given with no extra retroaction, while Fig. 14 shows the effect of retroaction. In this case a much greater increase of amplification is possible by the use of retroaction, but the best performance is only slightly better than in the medium wavelength case. This effect is due, no doubt, to the much smaller inherent retroaction at the

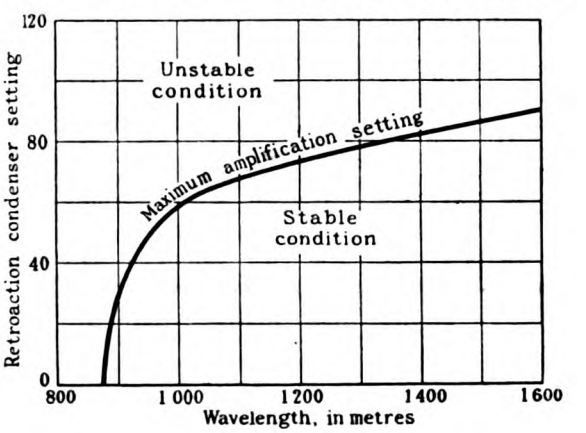

FIG. 15.

longer wavelengths, the curves for "no retroaction" showing a much reduced amplification. Again, the rapid rise in amplification at the lower wavelength end and the instability coupled with it are to be noted. The critical retroaction setting is shown in Fig. 15, from which it will be noticed that the receiver can be rendered unstable at all wavelengths with a smaller

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retroaction setting than in the previous medium-wave case, but, as before, it is unstable below a critical wavelength. In both cases the neutrodyne condenser was set to give balance by the usual audibility method, but

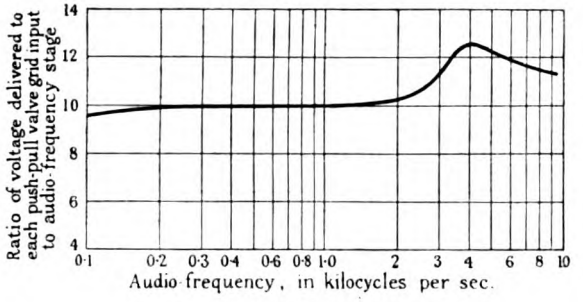

FIG. 16.

no guarantee can be given that this setting was absolutely exact.

It is to be noticed that several characteristics show output voltages of 40 or so. It is to be realized that, where such points occur, they are obtained by calculation only. For instance, if a characteristic is being obtained on the basis of a constant input of 5 mV and an output of 2 volts is obtained at a certain wavelength, another wavelength may require only 1 mV to give the same output. Making use of the known overall properties, the output is then computed in terms of a 5-mV signal. Such computation is permissible and renders the curves more useful, but at no time during the tests did the output exceed the limit of linearity of the output stage characteristic.

The last test performed upon this receiver consisted in maintaining a constant input and modulation while varying the modulation frequency. In this way the audio-frequency characteristic of the amplifier stage preceding the output valves was obtained and is shown in Fig. 16.

(*c*) *A Commercial Short-Wave Receiver* (15–60 *m*).—To examine the utility of the apparatus at short waves, an examination of a standard short-wave receiver employing one stage of radio-frequency amplification was undertaken.

The receiver circuit diagram is given in Fig. 17. There is one stage of radio-frequency amplification, the valve used having a voltage factor of 13 with an internal resistance of 30 000 ohms, followed by a rectifier valve with the same voltage factor and internal resistance, and lastly any normal output valve. Since the receiver is designed for use with weak telephonic outputs, the valve voltmeter was connected directly across the telephones and the power stage was calibrated with this connection. The known input voltage was supplied to either the input tuned circuit, the grid of the radio-frequency valve or the rectifier grid, by which means the performance of each part of the receiver could be analysed. Two tests were carried out at all wavelengths, namely, one with a variable modulation at a definite input, and the other with a variation of the input at a definite modulation. The modulation frequency was 1 000 cycles per sec. throughout, representative of an average audio-frequency in telephones with heterodyne reception of continuous waves.

The rectifier was calibrated at 30 and 60 m and found to be sensibly independent of frequency and to obey approximately a square-law relationship. By the

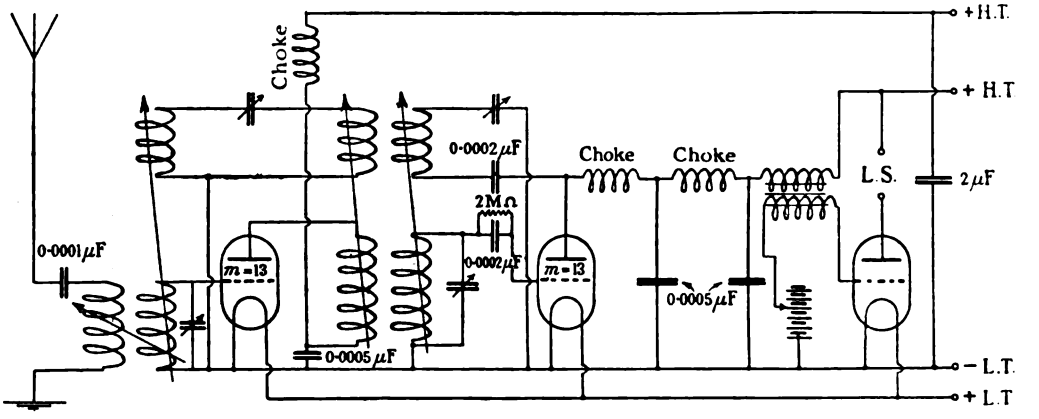

FIG. 17.—Circuit diagram of commercial short-wave receiver (15–60 metres).

use of this characteristic, together with those obtained by supplying the input tuned circuit or the radio-frequency stage, an analysis of the radio-frequency

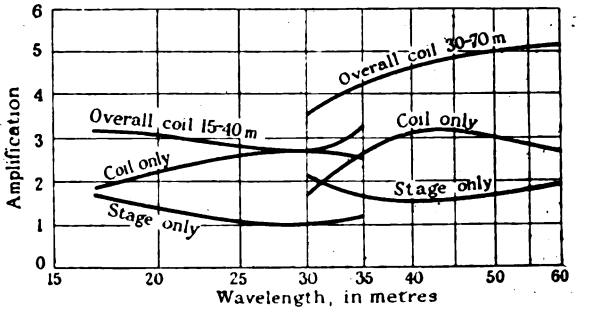

FIG. 18.

circuits was rendered possible. It is only possible to separate the coil and radio-frequency stage factors for the no-retroaction condition, since the amplification

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given when considerable retroaction is present is produced by the product of the two augmented values of both the coil factor and the radio-frequency stage factor. It is impossible to separate these two under critical conditions, owing to the modifying effect of the input resistance upon the circuits.

Within the limits of experimental accuracy, the output was found to obey a linear relationship with respect to the modulation percentage for constant input. With outputs greater than 0·3 volt on the power-valve grid the departures from this law were never greater than 5 per cent, and in the cases when the output was of the order of 0·1 volt they were much less.

It was found that the form of the input-output characteristic with constant modulation was independent of the frequency. For output voltages exceeding 0·1 this relationship was linear, but below this output voltage the characteristic was curved in a similar manner to the lower bend of a rectifier characteristic. The test made on the rectifier gave a somewhat similar curve, but the difference between the two was sufficient

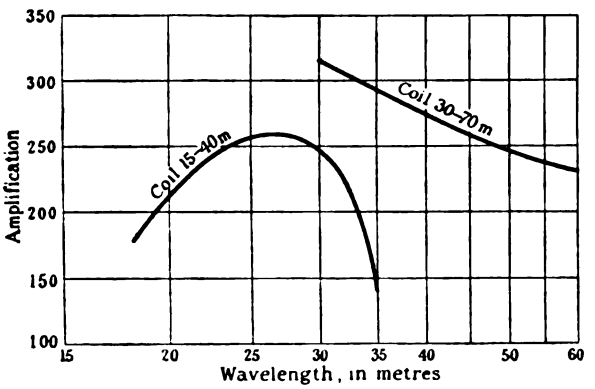

FIG. 19.

to suggest that a small amount of rectification was taking place in the radio-frequency stage. With critical retroaction the form of this input-output characteristic becomes saturated. This is due to the more critical retroaction setting possible with the weaker inputs, the larger inputs tending to throw the receiver more readily into oscillation. If the retroaction condenser had been adjusted for stability with a large input and left unaltered for the weak signals, this effect would have disappeared, but in the curves giving critical retroaction conditions the retroaction was adjusted for each measurement to give the greatest possible amplification. Owing to the difficulty of obtaining a critical setting, these characteristics have an accuracy of not more than 10 per cent.

The receiver is provided with two sets of coils to cover the wavelength band of 17–60 m. The overall radio-frequency amplification obtained by the input tuned circuit and the stage of amplification before the detector is shown in Fig. 18 with no retroaction, and the critical retroaction case is given in Fig. 19. In the former figure the overall amplification has been resolved into its two components, namely, the coil factor and the audio-frequency stage factor.

The general properties to be noticed by an examination of these results are that, owing to capacity shunting effects, the sensitivity falls as the frequency rises in the no-retroaction case; this is clearly shown in the case of the coil having the range 30–60 m (Fig. 18). The other effect of raising the frequency is to increase the retroaction, resulting in greater amplification. This is particularly noticeable below 30 m where the sensitivity is maintained. This characteristic agrees with the results obtained on the previous amplifier. In the critical retroaction case, where the input voltage is 1 mV the sensitivity increases as the wavelength is reduced, owing to the greater retroaction effect on the shorter waves, but it appears that at wavelengths below 25 m the retroaction condenser cannot be set so critically and a loss of sensitivity takes place.

In the critical retroaction case the measured performance given in Fig. 19 is valid only for inputs not exceeding 1 mV, for with greater inputs the amplification is found to be materially reduced. It would appear at

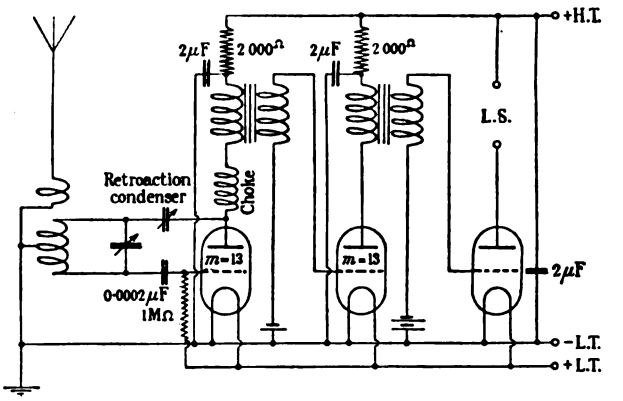

FIG. 20.—Circuit diagram of screened short-wave receiver (15–80 metres).

first sight that this behaviour is due to amplitude distortion, but on examination it is found to be due to the inability of obtaining so critical an adjustment of the retroaction condenser for the higher inputs. The results obtained on larger inputs suggest that the amplification with very weak inputs may be greater than that shown for an input of 1 mV in Fig. 19. The most striking feature of the results is the very poor amplification of the radio-frequency stage. This gives an amplification which never exceeds 2·1 and is only 1 at a wavelength of 28·5 m. It is interesting to note that the maximum coil factor is obtained at the same wavelength as the minimum radio-frequency stage factor for both ranges, and also that the maximum amplification obtained by the use of critical retroaction is obtained at the wavelength corresponding to the minimum radio-frequency stage factor. Since the use of retroaction is responsible for such a large increase in amplification, and since the radio-frequency stage appears to help to such a small extent, one arrives at the conclusion that methods of critical retroaction adjustment are more important than amplification at the high radio frequency by separate stages. Later experiments have confirmed this view. To illustrate

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this point more fully, the next test was made upon a short-wave receiver covering the same wavelength range but employing no radio-frequency amplification.

(*d*) *A Screened Short-Wave Receiver* (15–80 *m*).—The circuit arrangement of this receiver is shown in Fig. 20. Each stage was screened, and the coils used were made

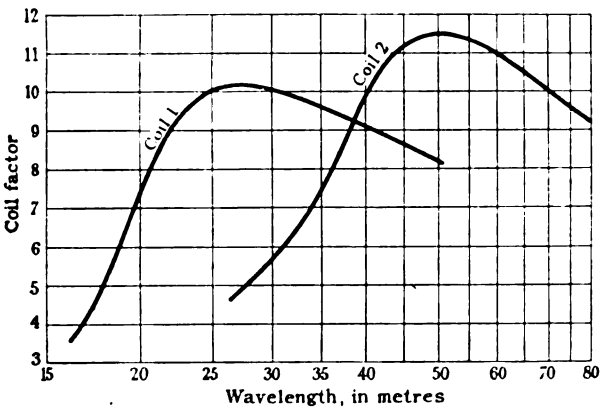

FIG. 21.

with bare copper wire well spaced, and with merely a skeleton frame to support them, so that the losses were reduced to a minimum. Two coils are required to cover the wavelength range from 15 to 80 m. It was found that this type of circuit was more easily adjusted than the previous one, largely owing to there being only two condensers, one of which controls the retroaction, whereas the previous circuit had three condensers, two of which were tuning controls.

By comparing these results with those of the previous receiver, Figs. 18 and 19, it is seen that a similar overall performance can be obtained by the use of a retroactive rectifier circuit, and that the predominant factor is the retroaction.

The radio-frequency stage in the previous amplifier was contributing but slightly to the overall result, and there appears to be little justification for the greater

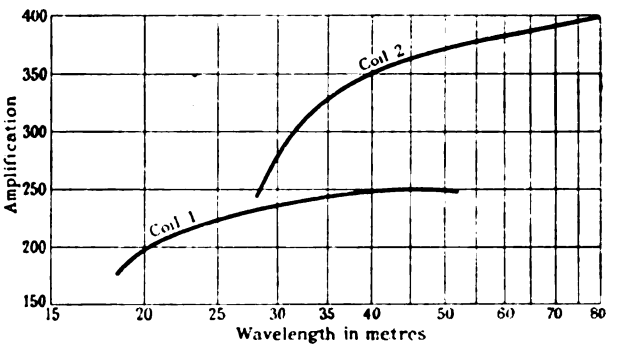

FIG. 22.

complexity of the controls and circuit involved by the use of a radio-frequency stage when a simple retroactive system can give the same performance.

(*e*) *A Supersonic-Heterodyne Receiver with Frame Aerial* (50–500 m).—Finally, a complete examination of the properties of a supersonic-heterodyne receiver was made, and the results are expressed in such a manner as to be of direct application to the practical use of such receivers.

The receiver used was partly built up from commercial

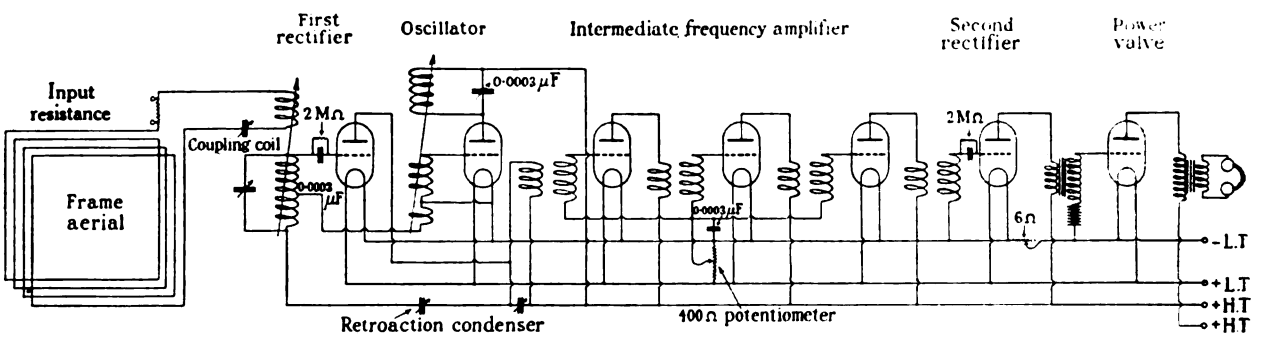

FIG. 23.—Circuit diagram of supersonic heterodyne receiver (50–5 000 metres).

By applying the known e.m.f. to the input tuned circuit and obtaining the audio-frequency output, and then by applying the e.m.f. directly to the rectifier, the coil factor can be obtained without deliberate retroaction. This characteristic is shown in Fig. 21, from which it will be noticed that the step-up due to the coil alone is much superior to that obtained in the previous case. This is largely due to the superior coils and also to the smaller load thrown upon the input circuit by the impedance of the rectifier valve circuit. By the use of retroaction, a greatly improved performance is possible, as shown in Fig. 22, the performance being similar with the two coils, namely, a fall in amplification with rise in radio frequency.

components, but was somewhat modified to suit the needs of a portable direction-finding receiver. The circuit arrangement is shown in Fig. 23. There is a separate oscillator having various coils to cover the required ranges, a 3-stage intermediate-frequency amplifier inductively coupled to the first rectifier, a second rectifier and an output stage. The voltage factors and internal resistances of the various valves are shown in Fig. 23.

The first characteristic to be obtained was the amplification of the three stages comprising the intermediate-frequency amplifier. This is given in Fig. 24, the potentiometer control of the grid bias to the second valve being so adjusted that the system was just stable

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and giving its maximum amplification. For this test the input resistance was connected directly to the grid of the first intermediate-frequency stage, and the characteristic shown in Fig. 24 represents the ratio of the voltage delivered to the second rectifier to the input voltage applied to the first intermediate-frequency stage. The type of characteristic obtained is a very general one for an intermediate-frequency amplifier and many similar characteristics have been obtained previously.* Owing to the inductive coupling between the first intermediate-frequency stage and the first rectifier, this latter characteristic does not represent the true selectivity of the intermediate-frequency section. To obtain this selectivity we must include the coupling between the first rectifier and the first intermediate-frequency stage. The input at the intermediate frequency was now applied to the first rectifier and a characteristic obtained between the voltage delivered to the second rectifier and the voltage applied to the first rectifier. Since the frequency of the applied voltage

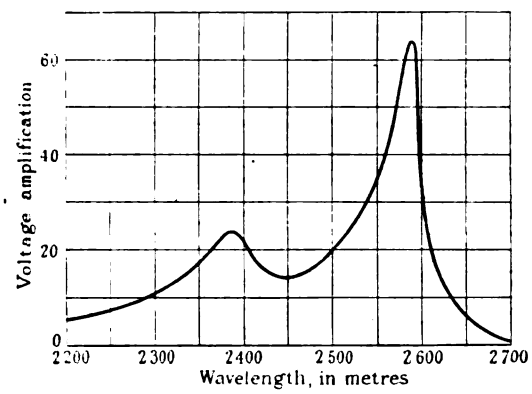

FIG. 24.

was that corresponding to the intermediate-frequency amplifier, such a test does not truly represent the conditions existing on the first rectifier, but it does give the true selectivity of the intermediate-frequency stages. This characteristic is shown in Fig. 25 and can be used to calculate the selectivity of the input circuit at any desired frequency.

Using this characteristic, the selectivity of the receiver has been calculated for the particular wavelength of 325·7 m and is plotted in Fig. 26. By applying the known modulated e.m.f. to the frame-aerial circuit, an experimental selectivity characteristic was obtained and is plotted alongside the theoretical curve for comparison. The agreement between the two curves is considered to be good enough for most general purposes.

The overall characteristic was now obtained. Tests were made to find the relation between the output audio-frequency voltage delivered to the grid of the power valve in terms of the voltage applied to the frame aerial by means of the screened resistance. This resistance had a value of 0·76 ohm, and consequently the effective resistance of the frame aerial was slightly increased by this amount. However, the error introduced by this factor is small. To cover the wavelength range of 50–500 m two different frame aerials were needed, together with various coils for tuning the frame circuit and the secondary circuit (see Fig. 23). The various oscillator coils were marked to signify the approximate range of wavelength as follows:—Ultra-short, 100–200, 250–800, and the coils are referred to

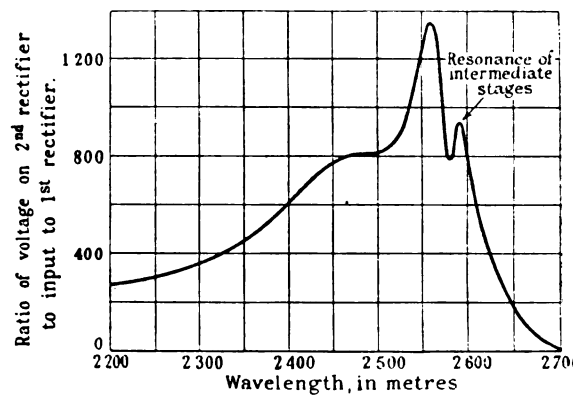

FIG. 25.

in the final results in this manner. Each wavelength can be received with two settings of the oscillator, one corresponding to the high setting of the oscillator condenser, and therefore to an oscillator frequency lower than that of the incoming signal, and the other corresponding to the low setting, in which case the oscillator frequency is higher than that of the incoming signal.

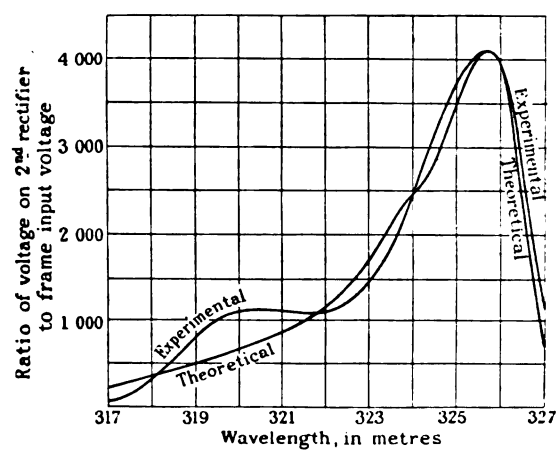

FIG. 26.

These two conditions are referred to as the " high " and " low " settings respectively.

For the purpose of expressing the results in the most useful form, a constant output signal has been taken, corresponding to average telephone strength. This was found to be given by a voltage of 3 mV on the grid of the power valve. Accordingly all the measurements have been referred to this value. In Fig. 27 the e.m.f. required in the frame aerial at a modulation percentage of 100 to give the standard output of 3 mV is plotted against the wavelength for the different ranges and

* H. A. THOMAS: "The Performance of Amplifiers," *Journal I.E.E.*, 1926, vol. 64, p. 253; and *Proceedings of the Wireless Section*, 1926, vol. 1, p. 12.

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oscillator coils. For other modulation percentages the output is found to be proportional to the modulation percentage and would thus be reduced proportionately. The large difference in sensitivity between the two oscillator settings is most marked, particularly at the longer wavelengths. The current in the oscillator was found to rise almost proportionately with the wavelength. With the long-range oscillator coil (250–800) the sensitivity is far greater on the low setting, i.e. with a weaker local heterodyne, and the shape of the curve conditions, above and below this optimum, represented by the long and short oscillator coils illustrate the utility of such characteristics as a guide to improving the design of a receiver. The difference between the sensitivity characteristics for the high and low oscillator settings is small at the short wavelengths owing to the small difference between the condenser settings and consequently the small difference in the strength of the local signal, whereas at higher wavelengths the difference is very large.

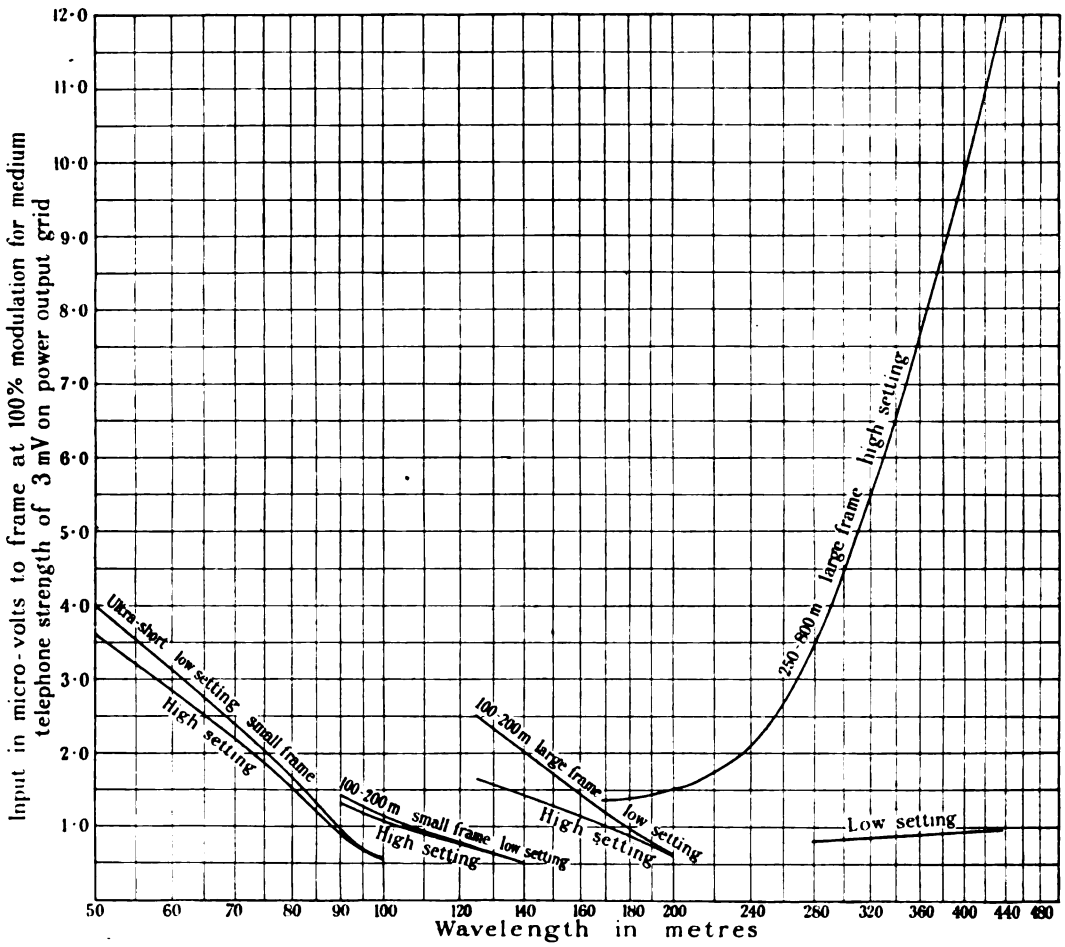

FIG. 27.

suggests that the sensitivity falls as the local heterodyne is increased with increase in wavelength. The obvious conclusion to be drawn from the form of this characteristic is that the local heterodyne is too strong. This could be improved by weakening the coupling to the frame aerial. On the other hand, the form of the remaining three sets of curves suggests that the local signal is too weak, since the maximum sensitivity always occurs when the oscillator wavelength is high, i.e. the local signal is large. It is well known that there is an optimum intensity for the local oscillation, and since no control is provided, it is not surprising that the optimum has not been obtained in this particular case. The two

By means of the known dimensions of the frame aerials, the potential gradient required to give the standard signal of 3 mV can be readily calculated. This has been done in Fig. 28. By the use of this characteristic, the field intensity (μV/m) required to give the standard signal can be obtained for any wavelength, and this characteristic represents the complete performance of the receiver.

SECTION (3).—SUGGESTIONS FOR SPECIFYING THE PERFORMANCE OF RECEIVERS.

From a perusal of the foregoing examples of the properties of receivers, it is apparent that the overall

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properties of any receiver can be expressed in terms of three characteristics:—

(*a*) The input voltage required at a definite modulation percentage to produce a definite standard output signal at all wavelengths within the designed range.

(*b*) The sensitivity variation in the neighbourhood of certain fixed wavelengths—this gives the selectivity of the receiver.

(*c*) The output variation with constant input and modulation percentage but variable modulation frequency—this gives the frequency distorting properties of the audio-frequency stages.

The limiting range of any receiver is easily obtained by this means, and the range for medium loud-speaker reception can also be obtained in a similar manner. In the case of an open antenna this latter characteristic involves a knowledge of its effective height. In view of the accuracy with which field-intensity measurements can now be made, and the consequent knowledge of the field distribution from a transmitting station, the author feels that if similar field-intensity/output characteristics were forthcoming for receivers, a considerable step would be taken towards more scientific classification of types, and the immediate result of such a classification would be to improve the design of radio receivers.

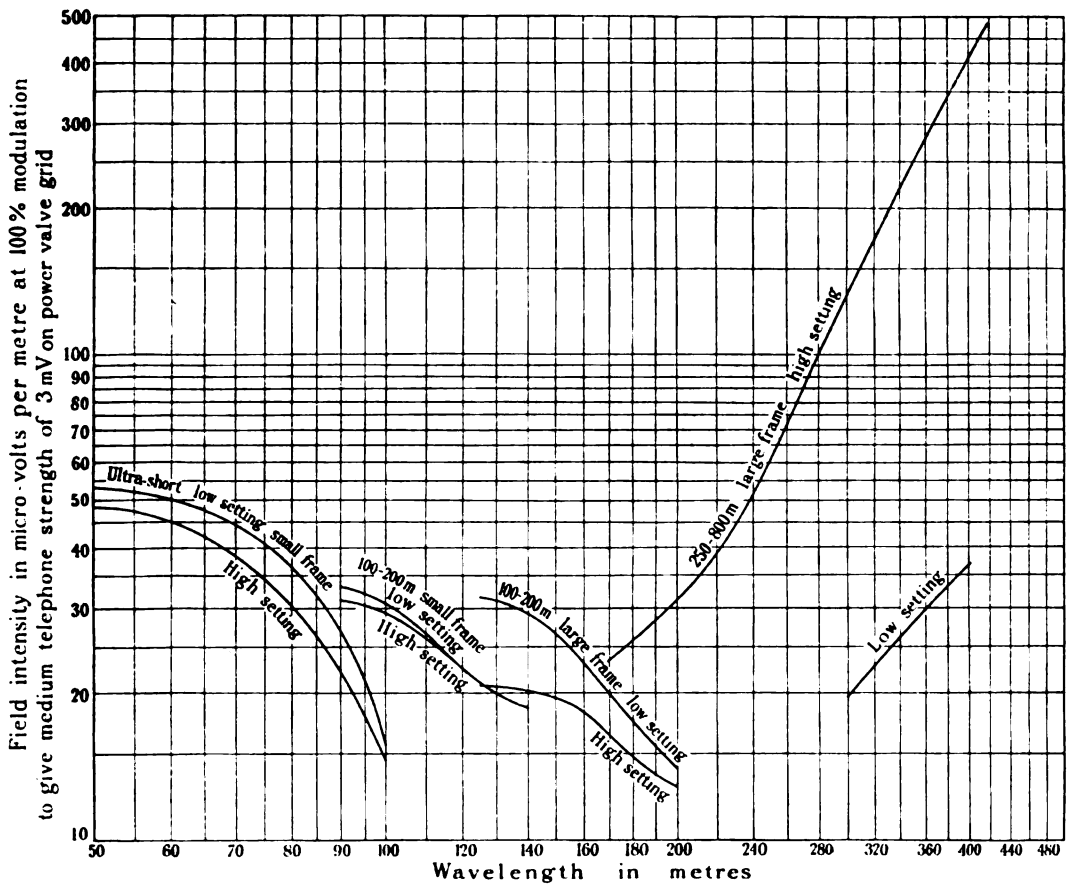

FIG. 28.

These three characteristics completely define the overall properties and, by means of them, the utility of a particular receiver can be obtained for any particular function. Oakley * has similarly concluded that these three properties are required to specify the overall performance of a receiver.

To determine the range of a receiver with respect to a particular transmitting station, two characteristics are required:

(i) The field-intensity characteristic of the transmitter expressed in the form of contour lines on a map.

(ii) The field-intensity/output characteristic of the receiver.

* *Loc. cit.*

In a somewhat similar manner the classification of receivers into various selectivity categories would help towards the correct selection of a receiver for a particular purpose. The selective properties of a receiver could be expressed as a figure which could be appended to the previous characteristic at the different wavelengths covered by that characteristic. Such a simple system would simplify the final statement of performance and would, in fact, enable the radio-frequency properties to be expressed in the form of a single curve.

The audio-frequency properties of a receiver have received considerable attention, chiefly because such results are comparatively simple to obtain, and un-

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doubtedly the progress in the design of audio-frequency amplifiers has followed measurements which have revealed the faults in the older types of amplifier. By means of an overall measurement, however, this characteristic would form a part of the tests and would take its place with the complete statement of the performance.

This work was carried out as part of the programme of the Radio Research Board and is communicated by permission of the Department of Scientific and Industrial Research. The author wishes to acknowledge the assistance given him by Dr. R. L. Smith-Rose and Mr. F. M. Colebrook, and also by his assistant, Mr. C. W. Spencer, in the latter portion of the work.

Discussion before the Wireless Section, 15th January, 1930.

Dr. J. Hollingworth: In considering such a comprehensive piece of work as that described in this paper I think it is right that some recognition should be made of the special problems which have had to be dealt with. In a subject which is developing as rapidly as radio-telegraphy, one of the chief difficulties is that developments on the practical side tend to take place so rapidly that the scientific side has difficulty in keeping pace with them. In this particular case, when the research first started about six years ago bright-emitter valves with factors of 2 or 3 were still in general use and overall amplifications were comparatively small. With the development of valves of ever-increasing factors, however, continual re-design of the apparatus was needed to cope with the increasing sensitivity and accuracy needed. Another factor which invariably occurs in developing apparatus for a measurement which has not reached the stage of a routine technique is the difficulty of proving that the apparatus really measures the quantity it is designed to measure. It is, of course, impossible to determine this by a direct method and it can only be decided by a long and difficult process of cross-checking. The paper also emphasizes a most important point which, though generally accepted now, had a long fight for recognition, namely that a receiver-amplifier is fundamentally a single unit from aerial to loud-speaker, and except for pure theoretical study cannot be treated as an aggregate of parts. On the instrumental side two small points arise, both in connection with the screened resistances. The first is that the intervals of resistance given on page 83 seem to be rather irregular; one would have expected the interval given as 0·96 ohm to be nearer 2 ohms to give a regular series. The other is that these resistances, although absolutely suitable for their work, are rather elaborate to construct. In the light of experience gained with their use can the author say whether any form of simpler and less expensive construction could be used without sacrifice of accuracy? The remark on page 85 with regard to the loads placed on a circuit by a valve voltmeter may surprise many, but is not altogether unexpected by those of us who know from practical experience how extremely sensitive an amplifier may be to alterations in the output circuit. With regard to the statement on page 86, " It has been found that the output current given by the radio-frequency oscillator is sensibly proportional to $(V - K)$, where V is the high-tension d.c. voltage and K is an experimental constant," do not variations sometimes occur in the value of K? Some time ago, when valves were not so reliable, I used to find that K varied from day to day. With regard to the actual amplifiers themselves, the results completely confirm my experience that on short waves, retroaction is the vital factor. In a paper by myself and Mr. Naismith, read before this Section in 1929 * we described the application of this idea to the use of an untuned aerial. The hope which we then held, viz. that we were not losing a great deal, seems now to be justified. In my opinion it pays to sacrifice to a limited extent the amplification of the set in order to ensure optimum conditions for the retroaction. In the author's opinion, what is the probability, in the average set, of accurately repeating a reading? Although at one time such an achievement was regarded as impossible with an amplifier, this should not be the case provided that the retroaction is carefully designed.

Mr. P. K. Turner: I should like to know whether there is any means of communication between the two cabins A and B. The point is important, because if, for example, a telephone is connected between them the screening will probably be adversely affected. In the first paragraph the author states that the most important measurement which can be made in connection with a receiver consists of the characteristic relation between input and output, together with an analysis of the behaviour of the individual portions of the apparatus. Does not such a measurement embrace all the information which is obtainable in connection with the receiver? In connection with the statement that the whole of the screening was done with iron sheet, I should like to know whether this material was employed merely because it was found to be satisfactory, or whether it was considered to be definitely superior to copper in its screening effect. With regard to the point which Dr. Hollingworth has also mentioned, in connection with the heavy load imposed by a valve voltmeter, I have used a number of valve voltmeters under many varying conditions and have never encountered any serious trouble from the load at audio frequencies,. A valve voltmeter which produces an apparent saturation curve of the type shown in Fig. 4, is open to question, and I should like further information about it. In measuring up to 40 or 50 volts with such instruments I find the load imposed by the voltmeter to be that of a small condenser with a power factor of the order of 0·1 and a capacity of 12 to 20 μF. These values are quite definite, and, within the limits for which the voltmeter is designed, are independent of the voltage actually being measured. The type of receiver used for broadcast reception is of greatest interest to me. Most of the measurements made by the author are not of importance in connection with instruments of that type. For example, the measurements given on page 85, where the modulation percentage is never

* " Portable Radio Intensity-Measuring Apparatus," *Journal I.E.E.*, 1929, vol. 67, p. 1033, and *Proceedings of the Wireless Section*, 1929, vol. 4, p. 140.

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greater than 15, are not of much interest to me. In the general circuit diagram of Fig. 5 and elsewhere there are indications that the output is being measured across a 100-ohm resistance in series with the loud-speaker. Thus the reading of the output is a function of the loud-speaker impedance, and as in many cases the set has to be considered apart from the loud-speaker it may, in those cases, considerably falsify the results. Such would be the case in connection with Fig. 16, which shows the audio-frequency response characteristic. Unless the loud-speaker and set are designed to be invariably used in conjunction one with the other, it is not justifiable to measure their performance together. An alternative method is to put a pure resistance load on the last valve, usually of about twice the impedance of the latter, and to measure the amplification of the set under these conditions. By a simple calculation the change introduced by the loud-speaker can be found. In Fig. 6 the screened resistance is shown on the grid side of the tuned circuit; if the resistance, one side of which is earthed, were, in fact, to be inserted at that point, no measurements upon the receiver would have any meaning, because of the alteration to the input circuit of the set. Should not the resistance be connected on the earthed side of the input circuit? Critical retroaction is usually understood to be the condition which approximates most closely to oscillation. This definition is not, in my opinion, altogether satisfactory, and I would therefore ask the author how it is possible to ascertain whether or not a system is in a state of critical retroaction. In cases where the retroactive arrangements are bad, the voltage is fed back several degrees out of phase with the input voltage, and critical retroaction is easily obtained. The output rises to a maximum and then diminishes slightly just before oscillation occurs. If, however, the retroactive arrangements are well designed the voltage fed back is almost in phase with the input, the critical stage before oscillation is very indeterminate, and the degree of amplification is closely dependent upon the precise setting. On page 88 the relationship between the output and the modulation percentage is mentioned. This is of first-rate importance, and I am sorry that no curves are given for varying modulation percentage at constant carrier voltage, showing how nearly the rectification is linear. The author states on page 87 that "a typical modern 4-valve receiver" was examined, but I am doubtful whether this is an accurate description of the receiver in question.

Mr. W. J. Brown: On page 85 the author states:—"the amplification must be expressed in terms of the voltage output delivered to the grid of the power stage with respect to some input voltage." Overall measurements on the receiver should definitely be made behind the output stage, as Mr. Turner has already suggested. His idea of including in the output circuit a resistance equal to twice the valve impedance seems to be a good one. Many of the author's measurements have been made under varying conditions of input and output, so that it is necessary in some cases to take input and output characteristics of the detector valve in order to interpret the measurements for constant input or constant output. I should like to suggest that, when specifying the performance of a receiver, all tests should be taken at a constant output power, and not necessarily with constant grid voltage on the power valve. The standards laid down by the Institute of Radio Engineers of America state that the measurements shall be made with a constant output of 50 milliwatts into an appropriate impedance. The sensitivity of the receiver is then expressed as the input signal which will give this constant output. A fixed percentage modulation of 400 cycles (40 per cent) is used for the purpose, while the input signal is usually expressed in terms of the field intensity in microvolts per metre, assuming a 4-metre aerial. In selectivity measurements the same principle of constant output might be employed in defining a standard test. The selectivity would then be measured by detuning the receiver by a certain amount and then increasing the input signal until the standard output had been obtained. The selectivity could then be expressed in terms of the detuning necessary when the input signal had been increased 10 or 100 times, for example. I am strongly in favour of the standardization of some form of test specification of the type suggested on page 95. I should like to ask the author how he has disposed of the difficulties to be met with in making measurements on a receiver which is close to oscillation. I suggest that it is almost impossible to measure, for instance, the selectivity of a receiver when the reaction is closely approaching its critical value. It may, therefore, be necessary to agree that such measurements should be made with the reaction adjusted to give, say, half the maximum attainable sensitivity.

Mr. T. H. Kinman: The measurements obtained on the higher frequencies are of particular interest to me and I should like to ask whether the inductive reactance of the input resistance, which is, of course, common to both filter circuit and input circuit of the receiver under test, was sufficient to cause interaction between these circuits when tuned to 10 megacycles and above. Such interaction would give rise to a variable reading on the input thermal meter "Q," Fig. 5. The author questions the value of the triode as a high-frequency amplifier for these higher frequencies. Has he made any measurements using a screen-grid valve for this purpose? My experience has been that with this type of valve not only is there a very useful high-frequency gain, but the tuning operation is, if anything, simpler, owing to the more uniform and less critical reaction adjustment required. No mention regarding the design of the thermo-couple is made in the paper, except that it was of the non-contact type. A thermo-couple designed for high-frequency measurements has been developed in America. It is assembled in such a way that the shunting capacity between its extremities is negligible. When dealing with very high frequencies, the usual construction is unsuitable owing to the shunting capacity across the pinch through which all the wires are usually brought out to base terminals.

Mr. A. G. Warren: I have encountered some difficulty in attempting to deduce Fig. 11 from Figs. 6, 7, 8, 9, and 10, arising from the fact that the conditions under which Fig. 9 was obtained are not given. The paper would be improved if these particulars were inserted.

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In connection with the voltage measurement, can it be assumed that the voltage applied is equal to the product of the measured current and the resistance? Although this is usually the case, it may be incorrect when low-loss tuning coils are employed. I cannot agree that the three characteristics given on pages 94 and 95 for specifying the performance of receivers supply all the necessary information. The first deals with sensitivity which, in broadcast receivers, is not important as the signals received are usually strong. With regard to the third characteristic, which is of great importance, I would re-write the closing phrase in this way: "This gives the frequency-correcting possibilities of the audio-frequency stages." As has been already remarked, the set has to be designed with the loud-speaker, and even the best loud-speakers call for correction by adjustment of the audio-frequency characteristics of the set. A listener requires that his receiver shall be able to discriminate at all intensities between wood-wind and strings, and that the tone of a piano shall be correctly reproduced. These conditions are not, in my opinion, included in the three characteristics. The first condition may be put into a practical form as the measurement of the second harmonic at a standard modulation of, say, 50 per cent. With the parabolic rectification which the author shows, the second harmonic might be rather distressing. I cannot, without further consideration, suggest a test appropriate to my second condition, but doubtless a means of estimating the response of the set to transients could be devised.

Mr. M. G. Scroggie: The author describes in detail his screened cabin for the oscillator, and then goes on to state: "The amplifier, too, must be housed in a screened room to prevent induction from external sources." Surely the oscillator is an external source and therefore need not be screened, provided that the screening of the amplifier is satisfactory. It is common practice in America to have only one screened compartment in apparatus for measuring the overall performance of receivers. The oscillator is just outside the screened compartment containing the amplifier, and cannot affect the latter except through the chosen channel. In production tests it is found possible in a number of cases to dispense even with the screening of the amplifier. This can be effected by doubling the frequency of the oscillator so as to produce sufficient difference in tuning between the amplifier and the oscillator to prevent direct action between them. Although used only for comparatively inaccurate tests, the method should be valuable because, as Mr. Turner mentioned, the apparatus shown by the author is rather difficult to operate. It differs from accepted American practice in the method of measuring the high-frequency input to the amplifier. Rather than to measure directly the small current applied through the resistance might it not be more convenient to use a less delicate current-indicating instrument followed by a calibrated attenuator? The method of measuring the output, which necessitates a separate calibration of the power stage in each case, seems to me to be unsatisfactory. In my opinion a suitable type of valve voltmeter presents no difficulty when used in an audio-frequency circuit. The difficulty of defining critical retroaction seems to be the one outstanding problem in this class of measurement. In America the standard broadcast receiver uses no intentional reaction and is usually incapable of oscillation, so that the difficulty does not arise to the same extent. In broadcast practice, at any rate, the difficulty will probably disappear from this country also, owing to the tendency to abolish intentional reaction. It is unfortunate that some receivers of more modern design were not included as examples. The standard practice, at any rate in broadcast apparatus, is to use screen-grid valves. These would show a far closer approximation to the amplification one would expect and would give, with ordinary components, a stage amplification of the order of at least 30 to 40 on the usual broadcast wave-band. The effect shown in Fig. 11, where the amplification increases rapidly at the high-frequency end, which effect is clearly due to unintentional feed-back, would not occur to anything like the same extent under the conditions of standard broadcast practice. The curve in Fig. 8 does, however, show up in exaggerated form the effect which is obtained even with screened-valve amplifiers, owing to the use of tuning circuits in which the capacity is variable and the inductance is fixed. If a curve of amplification against frequency is drawn, either for the threshold of instability or for constant stability, it will fall towards the high-frequency end owing to the greater admittance of the anode-grid capacity of the valve, as indicated in Fig. 8. It is desirable for the tuning characteristics to be such that the actual amplification curve is one of constant stability and therefore parallel to the threshold curve and on the lower side of it. This is obtained if both capacity and inductance are varied in a definite relationship, but a very good approximation to this ideal is obtained if the capacity is fixed. The amplification curve in the case of a variable-capacity tuning circuit cuts across the ideal curve almost at right angles. The result is that the amplifier tends to be too unstable at the high-frequency end and too "dead" at low frequencies. One advantage of a screened-valve amplifier is that the constant-stability curve is far more horizontal than that, for example, of Fig. 8.

Mr. J. H. Reyner (*communicated*): It is clear from an examination of the paper that a great deal of time and trouble have been involved in the preparation of the apparatus described. The author dismisses in a few words difficulties which must have taken weeks to overcome, and those conversant with methods of testing receivers must feel some disappointment that these difficulties have not been dealt with more fully. Previous speakers have commented upon the fact that the apparatus is of a precise nature and unsuitable for commercial use, although much of the data will be of value in the preparation of simpler equipment. It will be interesting to have the author's views on the possibility of producing a less complicated apparatus. I have been experimenting for some time with the production of small modulated high-frequency voltages which can be introduced into a receiver for the purpose of checking the performance. The voltage has been introduced into a dummy aerial or directly induced in the frame aerial of the receiver, according to circumstances. This is the American practice and does not involve any alteration to the receiver itself. It has been found pos-

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sible with a small apparatus some 2 ft. long, doubly screened, to produce and control voltages as small as 0·1 millivolt, which corresponds to a signal strength of 0·025 millivolt per metre, assuming a 4-metre aerial. This value of signal strength is distinctly smaller than that of the customary foreign stations which may be tuned in on any good modern set. It should be added that the experiments have been concerned mainly with frequencies below 1 500 kilocycles per sec., but some work has been done on much higher frequencies, with fair success. The most difficult part has been the production and calibration of a suitable method of coupling the receiver to the oscillator. The method which appears to be most satisfactory is an inductively coupled system which is calibrated by direct measurement, using a primary current many times larger than that employed in practice. This involves the assumption that the mutual inductance remains the same for small currents, which seems a reasonable supposition. The author condemns the inductively coupled method early in the paper, but it would appear to be the only practical method for a commercial testing equipment, since in order to obtain voltages as low as 0·1 millivolt by the voltage-drop system described in the paper the current would have to be too small to be measurable on any commercial instrument. An alternative method is the use of an attenuator, as mentioned by Mr. Scroggie. In this connection the Standardization Report for 1928 of the Institute of Radio Engineers is of considerable interest. Regarding the proposed specifications mentioned in Section (3), there can be no doubt that some such characteristic should be measured definitely on receiving equipment in order to obtain figures of merit. There is, however, a fourth characteristic about which we seem to know very little. This concerns the question of proportionality of input and output strengths. If the modulation of the incoming signal is increased by a certain amount, the low-frequency output should increase in strict proportionality, which it does not do in the majority of present-day receivers. The question is entirely separate from considerations of frequency distortion and is responsible for the poor behaviour of the average reproducing system on such instruments as the piano and similar instruments having a high percentage of transients. Mr. Warren has referred to this defect, but I believe that the root cause of the unsatisfactory reproduction is not well known. My experiments for some time past have all pointed to the lack of truly proportional amplification on differing strengths of signal as the fundamental defect. The greatest difficulty in the way of all such measurements arises from the use of reaction, which is still very popular in this country and is likely to remain so for many years. Not only does this affect the first of the proposed standard tests, but the second and third are also bound up with the amount of regeneration employed, while Mr. Turner has pointed out the extreme difficulty of defining the oscillation point. The attitude which I have continually encountered in discussions on receiver-testing is that while the performance of the receiver is dependent to such an extent upon the adjustment of the reaction control, any definite measurements are of academic interest only, and this point of view is difficult to combat. In America, simple modulated high-frequency oscillators are used as a matter of course by all concerns of any magnitude in the design, production and testing of their receivers. The technical public is, in consequence, educated to the idea and some real progress is being made in the matter. It would appear that the best method of procedure in this country is for the manufacturers themselves to commence forthwith to make receiver tests on the lines suggested in the paper. They will encounter difficulties peculiar to our own conditions and will doubtless devise their own methods of overcoming them. In due time, the industry as a whole will acquire the measurement habit which involves thinking and speaking in terms of definite data. If the present paper is responsible for development of this nature, the author will have rendered the industry a valuable service.

Mr. H. A. Thomas (*in reply*): With reference to Dr. Hollingworth's remarks, I would point out that many apparent defects in the final apparatus are due to the successive processes of rebuilding, modification and improvement which the equipment has undergone from time to time as points once considered essential have turned out to be of little account, whereas other details have been found to be all-important. All original apparatus is built in this way and, when in its final form, it is usually found that a great simplification can be made. If the apparatus were to be rebuilt it would take a quite different form as a result of the experience obtained with the original model. The input resistances have values which are not very accurately progressive, this being due to the difficulty of obtaining accurate lengths of the small-gauge resistance wire. The difficulties encountered in connection with the instruments were considerable and, since the range obtained by the first set of resistances was found to be sufficient for the purposes in hand, no modifications were made. Although it would be possible to simplify the design of these resistances, it must be remembered that accurate centring of the wire and adequate insulating and screening of the tapping lead involve rather a delicate construction. Comment is made regarding the effect of the valve voltmeter as a load on the output circuit, and I am glad that Dr. Hollingworth appreciates the effects which can be produced by such small loads when successively passed through the earlier stages of the amplifier. I have found that the value of K remains sensibly constant for the dull-emitter, low-amplification-factor, low-resistance valves that have been used. Undoubtedly the setting of the retroaction control to give critical reaction is delicate and the difficulty of reproducing conditions considerable, but with great care it was found that these critical reaction figures could be repeated to within 5 per cent in most cases.

In reply to Mr. Turner's remarks, I would point out that telephonic communication is maintained between the two cabins by means of screened leads which traverse a filter circuit of such design that no appreciable radio-frequency current passes along the line. Iron was used as the screening metal because of its superiority over copper. I would refer Mr. Turner to a paper by Dr. Smith-Rose * in which it is shown that the penetration

* R. L. Smith-Rose: "On the Electromagnetic Screening of a Triode Oscillator," *Proceedings of the Physical Society of London*, 1922, vol. 34, p. 127.

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of the eddy currents into the conductor is governed by the well-known "skin-effect" formula; the latter therefore determines the fraction of the original magnetic field which penetrates to any depth into the conductor. The decay factor governing the amplitude of the radio-frequency current at a depth x below the surface of the metal is $\epsilon^{-2\pi\sqrt{[\mu f/\rho]}x}$, where μ and ρ are respectively the permeability and resistivity of the metal, and f is the frequency of the alternations. Using this formula, the reduction in the field strength can be calculated for a screen of any given thickness and it can easily be shown that at medium radio-frequencies iron is 5·5 times more effective than copper if its thickness is the same. It is also clear that the screening effect is greatest at very high frequencies and that at audio frequencies effective screening becomes increasingly difficult. The load imposed by the valve voltmeter is not large if the last stage only is considered, but the effect of a small load may be serious when this load reacts back through the earlier stages of the amplifier. These were the conditions which were obtained in Fig. 4, and in view of the results it was considered expedient to use the voltmeter in the manner finally adopted. The figure for modulation (taken as 15 per cent) is an average value for good telephony transmission. Admittedly the figure is often greatly exceeded, but this fact does not invalidate the contention that 15 per cent is considered to be sound practice. In any case, the results with a greater percentage of modulation can be readily deduced from the curves given in the paper, since proportionality was found to exist up to a modulation percentage of 60. Mr. Turner criticizes my method of measuring the output and suggests that the properties of the amplifier should be determined with a resistance load, from which the actual properties with a loud-speaker can be estimated by "a simple calculation." I disagree with this—a view which is only too common. If the performance is determined with such a resistance load the input impedance of the last stage is greatly modified, and this impedance reacts back successively through each stage to the input. The effect on the early stages may be very considerable, especially if these stages have a high voltage-factor, and the calculation of the performance with an inductive load instead of a resistance load is far from simple. Again, the loud-speaker impedance is a very complex quantity involving a variable static and dynamic resistance and a static and dynamic reactance at different frequencies. In view of this fact it is essential that the receiver be measured under the actual conditions of use and I have laid great stress upon this point in the paper. Mr. Turner will see that I calibrate the last stage by applying to the grid an audio-frequency e.m.f. of the same frequency as the modulation, and obtain readings on the valve voltmeter. The 100-ohm shunt of the latter produces a negligible series modifying effect, and by this method the actual load is always in circuit. I agree that it would be an advantage to have a definite figure of amplification for the amplifier irrespective of the load, but this cannot be obtained, since every change in load affects the amplification characteristic. I must apologize for an error in drawing Fig. 6. The screened input-resistance is always used on the earthed side. My reply to Dr. Hollingworth covers the next portion of Mr. Turner's remarks on critical retro-action. A maximum figure was felt to be necessary and, as previously stated, it was found possible to obtain this with sufficient accuracy for most purposes.

My reply to Mr. Turner covers the first part of Mr. Brown's comments. I cannot see that any advantage is to be gained by substituting a power relationship. It appears to me that an amplifier is essentially a voltage device and that, having obtained the voltage on the power-valve grid, it is easily possible to obtain the output power if the last stage is considered to be a power relay and the input is regarded as a controlling voltage. The American tests give selectivity characteristics in terms of the input signal at the various radio frequencies necessary to provide a constant output; these characteristics have the appearance of an inverted resonance-curve. Either method of expressing selectivity is permissible, but I do not see that much is to be gained by departing from well-established practice in this matter. The appearance of these characteristics is puzzling at first. The suggestion that the reaction should be so set that the amplification is one-half of the maximum attainable is interesting, but does this not involve a previous determination of the maximum value?

With reference to Mr. Kinman's remarks, the reactance of the input resistance was always so small that its effect was negligible up to a frequency of 10 000 kilocycles per sec., and at 20 000 kilocycles it only added about 5 per cent to the net impedance of the input resistance. The tests were performed before the screen-grid valve was manufactured commercially, and consequently the remarks made in the paper are not meant to apply to such valve circuits. The thermo-junction used was of a well-known standard commercial type. It is interesting to note that recent work by Mr. F. M. Colebrook at the National Physical Laboratory has confirmed the fact that this type of junction is suitable for current measurement up to frequencies of the order of 10 000 kilocycles per sec., with an error not exceeding a few per cent.

I must thank Mr. Warren for pointing out the absence of sufficient information in Fig. 9. I have inserted the necessary information for the *Journal*. The point raised regarding the value of the voltage inserted in the tuned circuit is important. I agree that if the tuned circuit has a very low resistance it will exert an appreciable shunting effect upon the input resistance. In the tests described, however, the magnitude of this effect was never in excess of 7 per cent. In the case of critical reaction the shunting effect may be serious, but difficulty arises as to the determination of the effective resistance of the input tuned-circuit. I agree with Mr. Warren that a test of the response of the audio-frequency portion of the amplifier to transients would be very valuable, but the technique of this problem is not at present available. The response to a sudden applied signal, such as a square wave-shape, would be an indication of the effects which would be produced if a transient were applied.

Mr. Scroggie refers to the use of two cabins: it would be possible to dispense with one of them in a reconstruction of the apparatus, but the remaining small cabin, which at present contains the permanent oscil-

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lators, would have to house the receiver—and the latter could not be erected in such a confined space without radical alterations in the construction of the cabin. It must be remembered that the apparatus described has been produced by repeated modifications and the final arrangement is, admittedly, not perfect. I do not agree that a larger current and an attenuator are preferable to the method described. In the first place, it is not true that the accuracy of measurement of a large radio-frequency current is superior to that of a small current. As has already been stated, recent work by Mr. Colebrook has shown that the small current measurements given by vacuum junctions are reliable to a few per cent up to very high frequencies of the order of 10^7 cycles per sec., and, even if attenuators were accurate, I see no necessity for using them if the ideally simple screened resistance can be employed. Again, the impedance of the secondary of the attenuator introduces an impedance into the input tuned circuit which must be allowed for, and the screening needs to be more elaborate in the case of the larger current.

In reply to Mr. Reyner, I fear that a detailed statement of all the difficulties encountered would make the paper unduly long and would make rather tedious reading. As I have already pointed out, the apparatus could undoubtedly be materially simplified if there were a demand for such a complete reconstruction. I do not think there is any great difficulty in obtaining an input of 0·1 mV by means of a screened resistance. H. D. Oakley * uses this method with satisfactory results. With reference to Mr. Reyner's concluding remarks, I hope that the manufacturers of this country will adopt statements of a more scientific nature regarding the performance of their receivers than they have done hitherto. The object of the paper is mainly to indicate the lines along which development in this direction should take place.

* H. D. Oakley: "Quantitative Determination of Radio Receiver Performance," *Transactions of the American I.E.E.*, 1927, vol. 46, p. 498.

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AN AMMETER FOR VERY HIGH FREQUENCIES.*

By Professor C. L. FORTESCUE, O.B.E., M.A., Member, and L. A. MOXON.

(*Paper first received 18th September, and in final form 20th December, 1929; read before the* WIRELESS SECTION *5th February, 1930.*)

SUMMARY.

This paper describes the construction and theory of a simple form of hot-wire ammeter intended for use at any frequency up to 100 million cycles per sec. A cylindrical form is adopted in order that the correction factors for various frequencies may be calculated with certainty, and figures are given for an instrument as actually constructed.

(1) INTRODUCTION.

An ammeter for measuring currents up to about 1·2 amperes at frequencies between 10^7 and 10^8 cycles per sec. has been described by E. B. Moullin.† It consists of two parallel conductors mounted on elastic supports, the deflection of these supports under the electrodynamic forces arising from the mutual action of the currents being used to measure the magnitude of the current. The corrections for the capacity effects and the distribution of the current over the cross-sections of the conductors are somewhat difficult to calculate. The impedance would appear to be rather high and is to some extent dependent upon the magnitude of the current being measured. Finally, the measurement of the movement of the conductors is made by means of a microscope, with the result that the instrument is rather expensive, inconvenient to use and by no means robust.

Several alternative proposals have been made. P. R. Coursey and others have proposed to use capacity dividers and thermo-junctions. Hot-wire ammeters of the ordinary type are commonly used in practice without correction. The measurement of the voltage across either a single-turn loop of wire or other calculable form of inductance by means of a rectifying valve has also been suggested.

The ammeter described in this paper is a hot-wire instrument so arranged that the correction can be calculated easily. The impedance is relatively low and the R.M.S. value of the current is given directly by the movement of a pointer over a scale.

(2) THE DIFFICULTIES OF MEASURING VERY HIGH-FREQUENCY CURRENTS.

These arise from:—

(*a*) The capacity effects in consequence of which the current in the measuring device is different from that entering the instrument.

(*b*) The high inductive reactance of even very short conductors, and the consequent difficulty of using parallel circuits.

(*c*) The disturbance of the circuit in which the measurement is being made, owing to the relatively high impedance of the measuring device.

With regard to (*a*), at a frequency of 10^8 the reactance of a capacity as low as 10 $\mu\mu$F is only 160 ohms, and where thorough screening is adopted—as it must always be—it is by no means easy to keep the parallel capacities down to so low a figure. Moreover, the distribution of the current along the conductors employed for its measurement is dependent upon the distributed inductance and capacity. It is not always permissible to assume that the distribution is uniform, and where a high degree of accuracy is necessary a correction must be made for this effect.

The high value of the inductive reactance makes it difficult to adjust any instrument depending upon the subdivision of the current in unequal arms. At a frequency of 10^8 the reactance of a lead having an inductance of only 0·1 μH is already 62·8 ohms. The

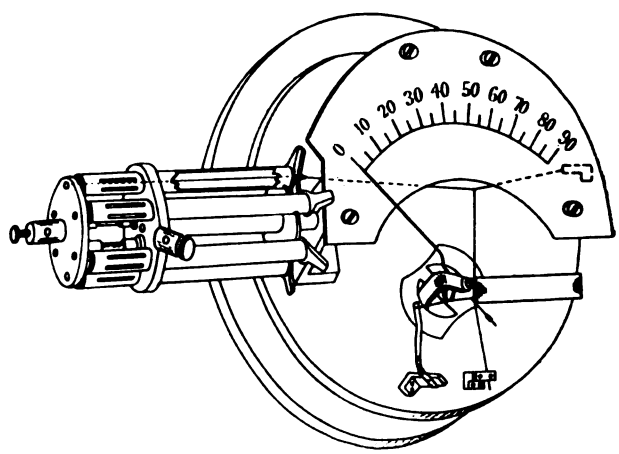

FIG. 1.

circuits must either be so adjusted that the impedances are identical, or the inductive reactance must be swamped by resistance or capacity reactance. In the latter cases the resultant instrument would have far too high an impedance for general use.

As to the permissible impedance, this depends upon the current for which the instrument is designed. The admittance should, if possible, be not less than 1 reciprocal ohm per ampere to be measured, since the total inductive reactance of resonant circuits at these high frequencies is usually low.

(3) PRINCIPLE OF THE NEW HOT-WIRE INSTRUMENT.

The simplest case for which all corrections can be applied with certainty is that of a wire of small diameter mounted co-axially with a cylindrical screen. Russell † has given a full treatment of this case. The high-

* Reprinted from *Journal I.E.E.*, 1930, vol. 68, p. 556.

† See page 133; also *Proceedings of the Royal Society*, A, 1928, vol. 121, p. 41.

† *Philosophical Magazine*, 1909, vol. 17, p. 524.

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frequency resistance of the wire and cylinder are easily calculable; the varying distribution of the current is easily found from the dimensions and a correction for the capacity of the terminal blocks can easily be made.

The passage of the current heats the wire and causes it to expand, and this expansion is used to measure the currents. The instrument is calibrated by direct current and the corrections for the high-frequency currents are calculated and applied either by reference to correction curves or by the use of curved scale divisions.

(4) The Construction of the Instrument.

This is shown in Figs. 1 and 2, where provision is made for six parallel wires each carrying a maximum direct current of about 1·2 amperes at full scale deflection.

Each current-carrying wire is kept in its central position on the axis of a screening tube by means of a flat spring at the end. One of these wires is directly attached to what is normally the heated wire of an ordinary hot-wire ammeter. As the wire in the tube expands and contracts, the pointer of the instrument moves to and fro over the scale and this movement thus becomes a measure of the current in the heating wires within the screening tubes.

A symmetrical arrangement of the terminal connections ensures uniform distribution of current between the heated wires.

For some experimental work this first instrument has been used with only one screened heating wire. This is quite satisfactory for the experiments, but the terminal capacity is much greater than need be the case with a simple wire, and consequently the corrections are abnormally large.

A second heating unit having only a single wire is now under construction. The terminal capacity is estimated to be only about one-fifth of that of the 6-wire heater, with the result that the corrections will be more normal.

(5) Calibration and Correction.

The initial calibration is by direct current in the ordinary way, and the scale is marked from this calibration.

The capacity between the terminal blocks and the screening tubes is measured before the wires are inserted. This and the dimensions enable the corrections to be made, the actual value of the alternating current entering the input terminal for each scale division being thus found and plotted for use as a correction curve. Alternatively, curved scale divisions as shown in Fig. 3 may be used.

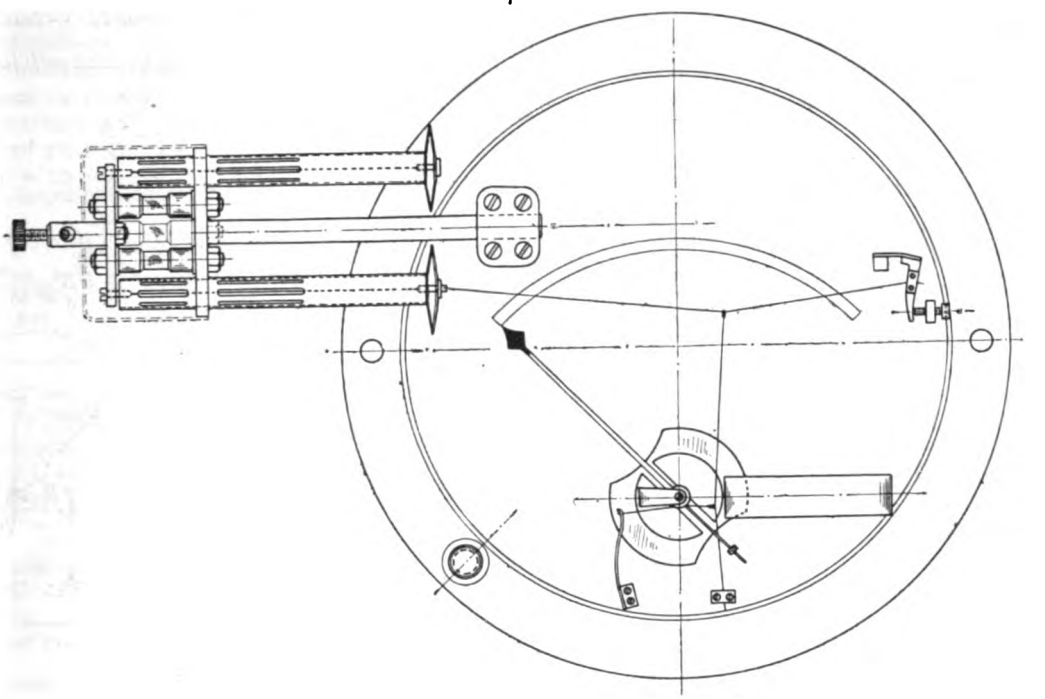

Fig. 2.—High-frequency hot-wire ammeter.
Scale: ⅔ full size.

(6) Calculations of the Corrections.

Notation:—

Terminal block capacity ..	C_t
Capacity per cm length of heating wire ..	C
Inductance per cm length of heating wire ..	L
High-frequency resistance per cm length of heating wire	R
Total current (R.M.S. amperes) entering input terminal	I_0
Current (R.M.S. amperes) in the terminal capacity	I_t
Current (instantaneous or vector amperes) in wire at distance x cm from input terminal ..	i
Ditto (maximum value)	$\mathfrak{I}$
Ditto (R.M.S. value)	I
Current (instantaneous, maximum and R.M.S. amperes) in wire at input-terminal end	i_w, $\mathfrak{I}_w$, I_w

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Voltage (instantaneous or vector volts) between wire and cylinder at distance x from input terminal v
Ditto (at input terminal) .. .. v_w
Number of parallel heating wires .. .. .. N
Length (cm) of each wire l

Considering firstly the distribution of current along the wire, the conditions are as usual; viz.

$$\left.\begin{aligned} -\frac{\partial i}{\partial x} &= C\frac{\partial v}{\partial t} \\ -\frac{\partial v}{\partial x} &= iR + L\frac{\partial i}{\partial t} \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad . \quad (1)$$

The terminal conditions are that when $x = 0$, $i = i_w$, and when $x = l$, $v = 0$, it being assumed that the screening tube is equivalent to a short-circuit. The methods given by Russell * must be adopted if this is not the case.

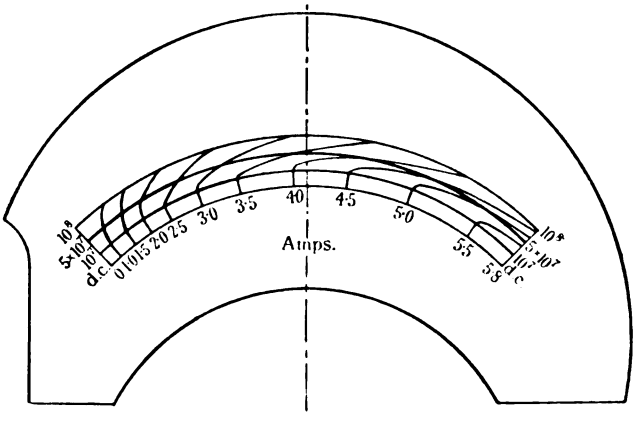

FIG. 3.—Scale for high-frequency ammeter.

Equation (1) and the terminal conditions give:—

$$i = i_w(\cosh mx - \tanh ml \sinh mx) \quad . \quad . \quad (2)$$

$$v_w = i_w\frac{m}{j\omega C}\tanh ml \quad . \quad . \quad . \quad . \quad (3)$$

where $$m = \alpha + j\beta = (-\omega^2LC + j\omega CR)^{\frac{1}{2}} \quad . \quad . \quad (4)$$

The mean power input is equal to the real component of the multiplier of i_w in (3) multiplied by I_w^2. If $I_{d.c.}$ is the calibrating direct current for the whole N wires, the current in one wire is $I_{d.c.}/N$ and the power expended is $(I_{d.c.}/N)^2lR_{d.c.}$, where $R_{d.c.}$ is the d.c. resistance per cm length of the wire.

If this power expenditure causes the same deflection as the alternating current, then

$$\frac{I_{d.c.}^2}{N^2}lR_{d.c.} = I_w^2\left(\text{Real component of }\frac{m\tanh ml}{j\omega C}\right) \quad (5)$$

Turning now to the current in the terminal capacity, this will be

$$i_t = i_w m\frac{C_t}{C}\tanh ml \quad . \quad . \quad . \quad . \quad . \quad (6)$$

The input current to the input terminal will be the sum of the capacity current given by (6) and N times the current in each wire; viz.

$$i_w\left(N + m\frac{C_t}{C}\tanh ml\right)$$

whence the value of the mean-square input current, I_0^2, can be found. Then, with the aid of (5), I_0^2 is connected with $I_{d.c.}^2$ and the correction factor determined.

From (4),

$$\alpha = \omega\sqrt{(LC)}\left[\frac{1}{2}\left\{\sqrt{\left(1 + \frac{R^2}{\omega^2L^2}\right)} - 1\right\}\right]^{\frac{1}{2}}$$

and

$$\beta = \omega\sqrt{(LC)}\left[\frac{1}{2}\left\{\sqrt{\left(1 + \frac{R^2}{\omega^2L^2}\right)} + 1\right\}\right]^{\frac{1}{2}}$$

When $\omega\sqrt{(LC)}$ is small, both α and β are small and

$$m\tanh ml = m^2l = (-\omega^2LC + j\omega CR)l$$

Also the real component of $m\tanh ml/(j\omega C)$ is then lR and, from (5),

$$I_w^2 = \frac{I_{d.c.}^2}{N^2}\cdot\frac{R_{d.c.}}{R} \quad . \quad . \quad . \quad . \quad . \quad (7)$$

With the same approximations, the input current is

$$i_0 = i_w\{(N - l\omega^2LC_t) + jlR\omega C_t\}$$

and

$$\begin{aligned} I_0^2 &= I_w^2\{(N - l\omega^2LC_t)^2 + l^2R^2\omega^2C_t^2\} \\ &= I_w^2(N - l\omega^2LC_t)^2 \end{aligned}$$

in most cases.

Substituting for I_w^2 in (7) gives

$$I_0^2 = I_{d.c.}^2\frac{R_{d.c.}}{R}\left(1 - \frac{l\omega^2LC_t}{N}\right)^2$$

or

$$I_0 = I_{d.c.}\left(1 - \frac{\omega^2lLC_t}{N}\right)\sqrt{\left(\frac{R_{d.c.}}{R}\right)} \quad . \quad . \quad (8)$$

The effective impedance of the instrument is

$$Z_0 = \frac{v_w}{i_0} = \frac{m\tanh ml}{j\omega(NC + mC_t\tanh ml)}$$

With small values of α and β this becomes

$$Z_0 = \frac{l(R + j\omega L)}{(N - l\omega^2LC_t) + jlR\omega C_t} \quad . \quad . \quad . \quad (9)$$

which at relatively low frequencies becomes

$$Z_0 = lR/N$$

as it should be for steady currents.

(7) CALCULATIONS FOR THE FIRST INSTRUMENT MADE.

Number of heating wires 6
Length of each wire .. .. .. 9·5 cm
Internal diameter of screening cylinder .. 0·80 cm
Diameter of wire .. .. 0·0140 cm
D.C. resistance of wire per cm length at 15° C. .. .. 0·1562 ohm
Specific resistance of wire at 15° C. $24·2 \times 10^{-6}$ ohm per cm per cm²
Observed terminal capacity .. .. 14·6 $\mu\mu$F

* *Loc. cit.*

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The capacity per cm length of wire is

$$\tfrac{1}{2}\log_e \frac{0{\cdot}80}{0{\cdot}0140} \text{ electrostatic units}$$
$$= 0{\cdot}1373 \times 10^{-12} \text{ farad.}$$

The inductance per cm length of wire is

$$2\log_e \frac{0{\cdot}80}{0{\cdot}0140} \text{ electromagnetic units}$$
$$= 8{\cdot}093 \times 10^{-9} \text{ henry.}$$

The ratio $R/R_{d.c.}$ when computed from the usual tables is as follows:—

Frequency ..	10^6	10^7	5×10^7	10^8
$R/R_{d.c.}$ ..	1·000	1·015	1·268	1·678

The approximations made in the previous section are permissible for all except the last case, which must be calculated from the complete formulæ.

The results are:—

Frequency	10^6	10^7	5×10^7	10^8
$\frac{I_0}{I_{d.c.}}$	1·000	0·992	0·874	0·706
Impedance of instrument, in ohms ..	0·261	0·840	4·11	8·84

(8) Comments on These Results.

(*a*) The corrections are very small for frequencies below 10^7, and from this frequency downwards the direct-current calibration can be used with errors not greater than 1 per cent. Above that frequency, however, the correction increases rapidly owing to the relatively quick increase of the resistance, and also because of the increasing current in the terminal capacity.

(*b*) The impedance of the instrument increases with the frequency, and the reactance is inductive up to frequencies considerably above 10^8.

(*c*) The magnitude of the correction due to the capacity is mainly dependent upon the term $\omega\sqrt{(LC)}$, which in this case is ω/c^2, where c is the ratio of the electromagnetic to the electrostatic absolute units of quantity. With the cylindrical construction, therefore, the dimensions will have little influence on this correction. There is consequently a fairly free choice of dimensions, which enables the design to be adapted to the other requirements of the hot-wire instrument.

(*d*) The results given in the previous section are only calculated to a moderate degree of accuracy because sufficient experience of either the construction or use of these instruments has not yet been obtained to justify more exact computations. These will be made later if it is found desirable.

(9) Programme of Further Work.

There is now in progress a lengthy research which includes both the design of the instrument, and the method of comparing the results with other similar ammeters or indirect methods of measuring current. This research is being carried out under the auspices of the Department of Scientific and Industrial Research, by whose permission this paper is now published.

The authors wish to thank Mr. D. W. Hopkin for assistance in preparing drawings for the illustrations.

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A METHOD OF MEASURING MECHANICAL IMPEDANCE.*

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(*Paper first received 2nd December, and in final form 20th December, 1929; read before the* WIRELESS SECTION *5th February, 1930.*)

SUMMARY.

The device the mechanical impedance of which at a given frequency is to be measured is driven by one prong of an electrically-driven tuning fork, and the other prong is loaded to restore balance, the load required being the mechanical impedance to be determined. The balance condition is indicated by the absence of vibration of the fork support.

An improvement depends upon the experimentally determined fact that mechanical resistance out-of-balance does not appreciably affect the support vibration. If the prongs are balanced for mass, the total resistance can be found from a decay-factor determination or by a single measurement of amplitude with a known current, and the resistance of the device estimated.

A measurement is made of the mechanical impedance at 100 cycles per sec. of a certain gramophone pick-up.

LIST OF SYMBOLS.

m = equivalent mass of each prong of a tuning fork (grammes).
s = equivalent stiffness of each prong of a tuning fork (dynes/cm).
r = equivalent mechanical resistance of each prong of a tuning fork (dynes/cm/sec.).
a = decay factor of each prong $\{= r/(2m)\}$.
m' = mass added to one prong of tuning fork (grammes).
r' = total mechanical resistance added to two prongs (dynes/cm/sec.).
r_1 = resultant mechanical resistance of each prong when loaded $(= r + \frac{1}{2}r')$.
f_0 = resonant frequency of tuning fork (cycles/sec.).
ω_0 = $2\pi f_0$ (radians/sec.).
C_0 = capacity (μF) in oscillatory circuit of valve oscillator to produce a frequency f_0.
x = amplitude of displacement of prong from mean position (cm).
u_m = maximum value of velocity of prong (cm/sec.).
x_0, u_0 } values of x and u_m at resonance.
i = R.M.S. value of alternating current in driving coil of fork (mA).
i_m = maximum value of alternating current in driving coil of fork (absolute electromagnetic units, called " abamps ").
A = force factor of driving system (dynes/abamp) $(= NB/S)$.
N = number of turns on driving coil.
B = flux density of permanent magnetism in the air-gaps (lines/cm^2).
S = total reluctance of the magnetic circuit of the driving system.

SECTION (1).

The resonant frequency of an electrically-driven tuning fork depends upon the length and thickness of the prongs and upon the material of which they are made, but referred to the point of drive an equivalent mass m and an equivalent stiffness s can be found for each prong, so that the resonant frequency is given by

$$f_0 = \frac{1}{2\pi}\sqrt{\left(\frac{s}{m}\right)} \quad . \quad . \quad . \quad . \quad . \quad (1)$$

This applies accurately for small amplitudes only. At large amplitudes the frequency of maximum response decreases as the amplitude is increased, according to a square law.†

The decay factor of each prong of the tuning fork can also be expressed in terms of an equivalent mechanical resistance r, as

$$a = \frac{r}{2m} \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

Here again only small amplitudes are intended, as the decay factor increases with amplitude.

If a mass m' is added to each prong at the point of drive the resonant frequency will be lowered to

$$f_0 = \frac{1}{2\pi}\sqrt{\left[\frac{s}{m+m'}\right]} \quad . \quad . \quad . \quad . \quad (3)$$

and if a mechanical resistance r' is added to each prong at the point of drive the decay factor will be altered to

$$a = \frac{r+r'}{2(m+m')} \quad . \quad . \quad . \quad . \quad . \quad (4)$$

If the two prongs of the fork are perfectly balanced there will be little or no tendency for the support to vibrate, but if they are not balanced the support will tend to vibrate in order that the centre line of the mass of the two prongs may remain stationary. Suppose now that one prong is made to drive a gramophone pick-up. Its constants will on that account be altered to $(m + m')$ and $(r + r')$, and the fork will be thrown out of balance and the support caused to vibrate. But if the other prong can be loaded with a known mass and a known resistance to restore the balance, the effective mass and resistance of the pick-up are determined. The case where the pick-up reactance is effectively a stiffness can be dealt with by loading each prong initially with a known mass m' sufficient to cause a greater reduction of frequency than the increase caused by the pick-up stiffness.

SECTION (2).

The obvious indication that balance has been obtained is the absence of vibration in the support, and many experiments were made in this direction.

The fork used was a laboratory model consisting of two bars of hard steel mounted on a bracket to form a

* Reprinted from *Journal I.E.E.*, 1930, vol. 68, p. 560.

† *Proceedings of the Physical Society of London*, 1926–7, vol. 39, p. 338.

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tuning fork. The effective length of the prongs could be varied by altering the position of a massive steel clamping block, the frequency range being from 80 to 500 cycles per sec. The fork was magnetized and between its prongs was mounted a small low-resistance coil with a core of a bundle of soft-iron wires to complete the magnetic circuit of the prongs.

In order to leave the support fairly free to vibrate, the bracket at the stationary end of the prongs was removed, the only support being a long bolt clamping the steel block to the base. The unused portion of the prongs was then available for mounting the vibration detector as in Fig. 1.

After trying several types of detectors the one found to be most useful was a set of 22 single reeds rigidly clamped into two brass blocks and of various lengths covering a frequency range of about 80 to 100 cycles per sec., the free ends being viewed from above through a low-power microscope. A button microphone and vibration galvanometer was much more sensitive but somewhat unstable, and the tuning of the galvanometer had to be adjusted with every alteration of balancing load, making the measurement very laborious.

The oscillator used had an L.S.6A output valve working from the 400-volt d.c. mains through a filter circuit, this large power output being necessary to drive the heavily loaded fork through a high-value swamping resistance necessary for stability.*

The balancing impedances used were (*a*) reactive, and (*b*) resistive, as follows:—

(*a*) A set of weights each shaped to clamp on to the prongs by means of a small set-screw.

(*b*) A 3-bladed copper paddle dipping into a bath of thick machine oil. The direction of motion of the paddles was in their own plane, this scheme allowing rigid construction consistent with lightness to be used.

The pick-up was driven from the fork by fixing a small piece of gramophone record to the top of the prongs by means of hot shellac and resting on this the needle of the pick-up. This coupling has the advantage of simplicity and, moreover, very nearly reproduces the forces between needle and record under actual playing conditions. Back-coupling due to vibrations passing along the pick-up arm and hence to the fork support was prevented by mounting the arm on a massive cast-iron base independent of the fork mounting.

The practical difficulties encountered with this apparatus are summarized below:—

(1) Resistance could not be added without also adding effective mass.

(2) Resistance out-of-balance produced an inappreciable vibration of the support.

(3) An out-of-balance of mass produced a considerable increase of effective resistance, which caused a decrease in the prong and support vibrations although balance conditions were being departed from.*

(4) Every change of added mass necessitated a change of frequency of drive. Unless the fork were driven at the resonant frequency corresponding to the load of the moment, it was impossible to say whether balance conditions were being approached.

It then appeared to be possible [from (2)] that with the order of resistances encountered the additional effective resistance of each prong would be one-half the total added resistance, irrespective of whether it was added to one prong only or divided between the two. In other words, a determination of the decay factor of the fork by any method,† provided of course that the mass of each prong was balanced, would give the resistance of the pick-up. In Appendix 2 it is shown that this idea is justified. The use of the support vibration as an indication in balancing the mass was possible but tedious for the reasons given above. A simpler method was found to be as follows: A curve of resonant frequency/added mass (equal mass to each prong) was drawn. Then the pick-up was added to one prong and mass added to the other prong until the resonant frequency of the fork for the mass added agreed with the value read from the curve. The effective mass of the pick-up was then equal to the mass added. Alternatively, equal known masses (greater than the pick-up mass) were added to each prong, and one of these was subsequently replaced by the pick-up and a variable mass, the latter being adjusted to bring the resonant frequency back to its previous value.

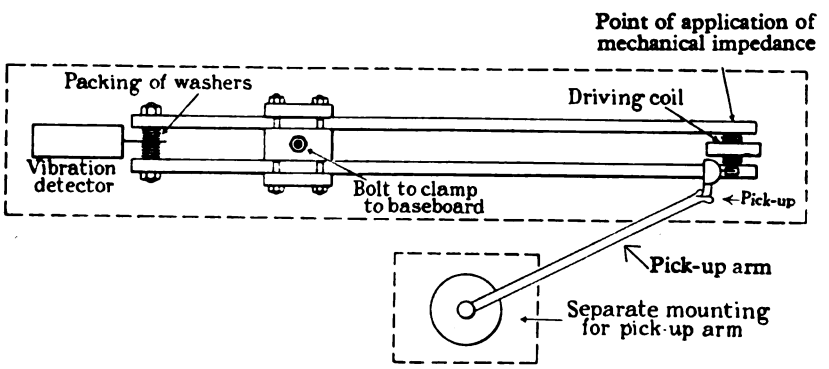

FIG. 1.

* *Proceedings of the Physical Society of London*, 1926–7, vol. 39, p. 33

* See Appendix 1. *Journal I.E.E.*, 1924, vol. 62, p. 517.

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SECTION (3).

Details will now be given of an experimental determination of the impedance of a certain gramophone pick-up.

The skeleton circuit used is given in Fig. 2. The fork amplitudes were measured by viewing some illuminated mercury globules on the prong tips through a high-power travelling microscope with a scale in the eye-piece, and the current was determined by the valve voltmeter and known resistance R_1, using a Weston relay as a critically damped reflecting galvanometer.* R_2 was a series controlling resistance for keeping the current constant while taking the resonance curve, the smallest value to which R_2 was reduced being sufficiently large to prevent unstable effects.

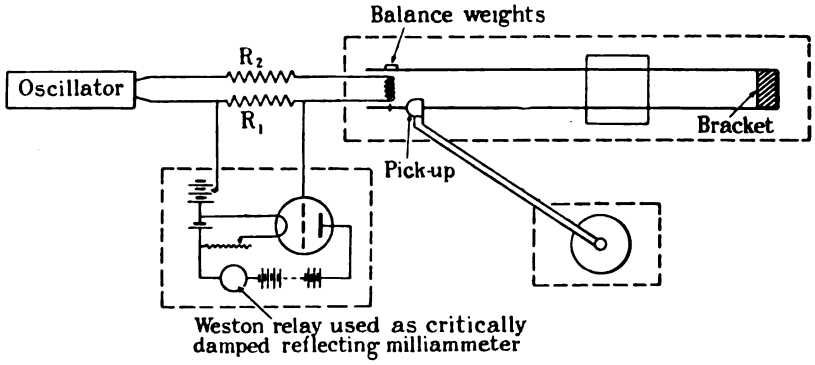

FIG. 2.

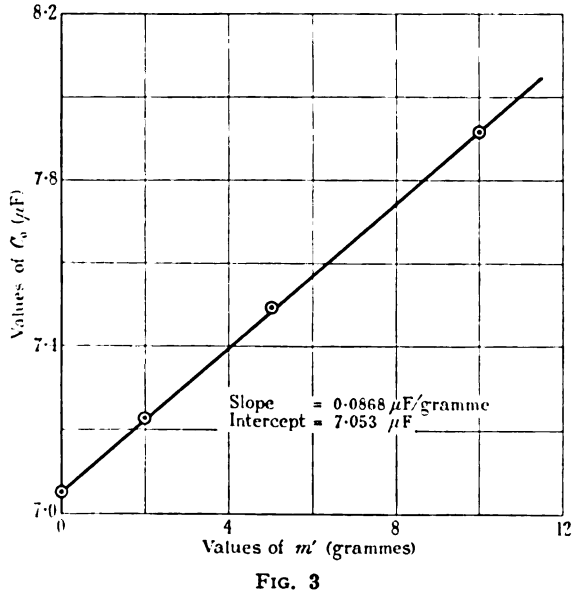

FIG. 3

The equivalent mass and resistance of the fork prong required in equations (3) and (4) were found by the following method:—

From (3) $$\frac{1}{f_0^2} = 4\pi^2\frac{m}{s} + 4\pi^2\frac{m'}{s} \quad . \quad . \quad . \quad . \quad (5)$$

and if C_0 is the capacity in the oscillatory circuit of the oscillator corresponding to a frequency f_0

$$f_0^2 C_0 = K$$

where K is some constant determined from the oscillator calibration.

Hence $$C_0 = 4\pi^2 K\frac{m}{s} + 4\pi^2 K\frac{m'}{s} \quad . \quad . \quad . \quad . \quad (6)$$

If, therefore, equal masses m' are added to the prongs at the point of drive, C_0 found for resonance, and m' plotted against C_0, the graph should be a straight line of slope $4\pi^2K/s$ and intercept $4\pi^2Km/s$, and from these two quantities m and s are determined.

This experiment was carried out, the graph obtained being shown in Fig. 3. This checks the theory and provides a rapid method of finding m and s for any fork.

K, from the oscillator calibration, was

$$10^5 \text{ (cycles/sec.}^2)^2\ \mu\text{F}$$

and from the graph the slope is 0·0868 μF/gramme and the intercept is 7·053 μF.

Hence $m = 81{\cdot}3$ grammes, and
$s = 45{\cdot}5 \times 10^6$ dynes/cm

A resonance curve of the unloaded fork at constant current was taken and by the circle construction the decay factor was found to be 0·096, giving

$$r = 2ma = 15{\cdot}6 \text{ dynes per cm per sec.}$$

where r is the mechanical resistance of one prong of the fork.

The pick-up was now coupled to one prong and at a frequency of 117 cycles per sec. it was found that the reactive load of the pick-up and 16 grammes was balanced by 4·3 grammes on the other prong.

Thus the pick-up has an apparent negative mass of $16 - 4{\cdot}3 = 11{\cdot}7$ grammes, corresponding to a stiffness, at this frequency, of

$$s_0 = \omega_0^2 m_0 = 4\pi^2 \times 117^2 \times 11{\cdot}7 \text{ dynes per cm}$$
$$= 6{\cdot}36 \times 10^6 \text{ dynes per cm}$$

A resonance curve taken of this balanced system gave

$$a = \frac{r + \frac{1}{2}r'}{2(m + m')} = 14{\cdot}3$$

hence $$r + \tfrac{1}{2}r' = 2\,450$$

giving $$r' = 4\,870 \text{ dynes per cm per sec.}$$

* *Journal I.E.E.*, 1925, vol. 63, p. 401; also *Proceedings of the Physical Society of London*, 1927-8, vol. 40, p. 14.

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In order to test the sensitivity of the method for the balancing of mechanical reactance, a graph was taken showing resonant capacity C_0 plotted against mass added to one prong only (Fig. 4). The steepness of the curve for small values of mass out-of-balance is very apparent, and a measurement can be obtained without any difficulty with an accuracy of 2 or 3 per cent.

The determination of the decay factor from the graphical construction, however, is laborious and not so accurate, and the advantages of a single-reading method for determining the resistance from the points of view of increased accuracy and rapidity of measurement are apparent.

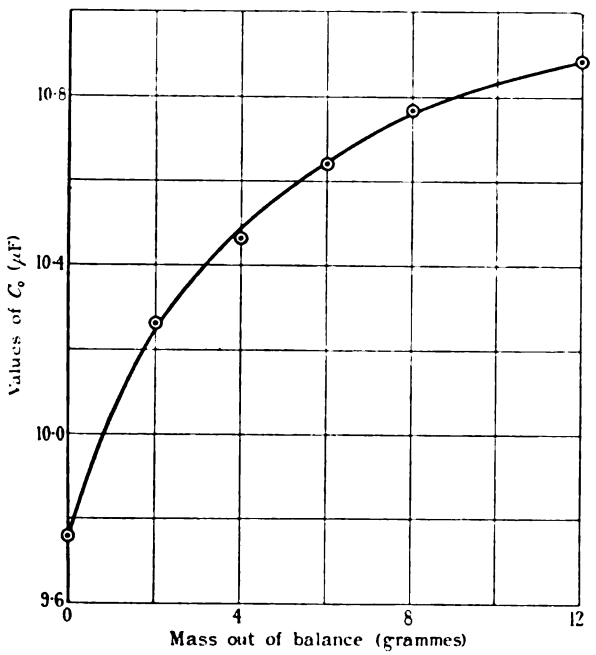

FIG. 4.

Such a method can be obtained by making use of the equation

$$u_0 = \frac{Ai_m}{r_1} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (7)$$

where x_0 = amplitude at resonance,
u_0 = max. velocity at resonance,
$= \omega_0 x_0$,
i_m = current in abamps (max. value),
A = force factor of driving system in dynes per abamp, and
r_1 = total mechanical resistance of one prong.

Hence if A is known, measurement of x_0, C_0 and i_m will determine the mechanical resistance $\frac{1}{2}r' + r$.

It is well known that

$$A = NB/S \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (8)$$

where N = number of turns on driving coil,
B = flux density, in lines per cm^2, of permanent magnetism in the air-gaps, and
S = reluctance of magnetic circuit,

so that from theoretical considerations A should be independent of the external load on the fork. The accuracy of this assumption is discussed in Appendix 4.

The experimental details for determining the force factor are given in Appendix 3, the result being $A = 5{\cdot}06 \times 10^6$ dynes per abamp.

The resonance curve giving the pick-up impedance was taken at a constant current of 26·0 mA and the maximum deflection at resonance was 0·0102 cm, so that, using the shorter method,

$$\tfrac{1}{2}r' + r = \frac{Ai_m}{\omega_0 x} = \frac{5{\cdot}06 \times 10^6 \times 26 \times 10^{-4} \times \sqrt{2}}{2\pi \times 117 \times 0{\cdot}0102}$$
$$= 2\,470 \text{ dynes per cm per sec.}$$

This gives

$$r' = 4\,910 \text{ dynes per cm per sec.}$$

which agrees with the result obtained by the graphical construction.

CONCLUSION.

The method has so far been tried only at the low frequency of about 100 cycles per sec., but no difficulty is anticipated in obtaining results at frequencies up to, say, 500 cycles per sec. with forks of higher pitch. It is hoped to extend the frequency range still further by employing the higher modes of vibration of the fork.

The experimental work described in this paper was carried out in the Telegraphy and Telephony Laboratories of the City and Guilds (Engineering) College, and the thanks of the authors are due to Mr. F. W. Andrews and the staff of the departmental workshop for assistance in the construction of the special apparatus used.

APPENDIX 1.

INCREASE OF EFFECTIVE MECHANICAL RESISTANCE OF FORK WITH A MASS OUT OF BALANCE.

In the first attempt made to measure mechanical impedance by a tuning-fork method,* the pick-up was driven by one prong of the fork, and the influence on the decay factor and frequency of the fork was used to estimate the mass and resistance of the pick-up. The mass was determined from a curve such as Fig. 4. To estimate the resistance, resonance curves were drawn with various masses added to one prong only of the fork and the decay factor as determined from these curves was plotted against the mass added, to obtain a curve such as Fig. 5. From this curve the increase in the effective mechanical resistance of the fork itself resulting from out-of-balance of mass can be estimated and subtracted from the total increase obtained with the fork driving the pick-up. To obtain an idea of the correction involved, reference may be made to Section (3). With the mass balanced the pick-up introduces a decay factor of 14·3, whereas the reactive load of the pick-up, being 11·7 grammes, would itself introduce, if unbalanced, a decay factor of 10·1.

This correction is of the same order as the resistance to be measured, and the scheme was therefore abandoned in favour of a balance method.

* The first experiments in which were carried out by Mr. R. Toombs.

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APPENDIX 2.

Constancy of Fork Resistance with an Unbalanced Resistive Load, and Measurement of Power Factor of Oil Damping.

For this experiment, matched, 3-bladed copper paddles could be attached to each prong at equal distances from the support, and glass containers so arranged that either or both paddles could be immersed to the same level in thick machine oil. By this means oil damping could be applied to either or both prongs while leaving balanced the reactive load due to the mass of the paddles.

The equivalent mass and stiffness of each prong with paddle attached was found as described in Section (3), and the decay factors under the various resistive loads were found by plotting resonance curves. With only one paddle immersed there was an out-of-balance of reactive load due to the oil carried along with the paddle. This was indicated by a change in the resonant

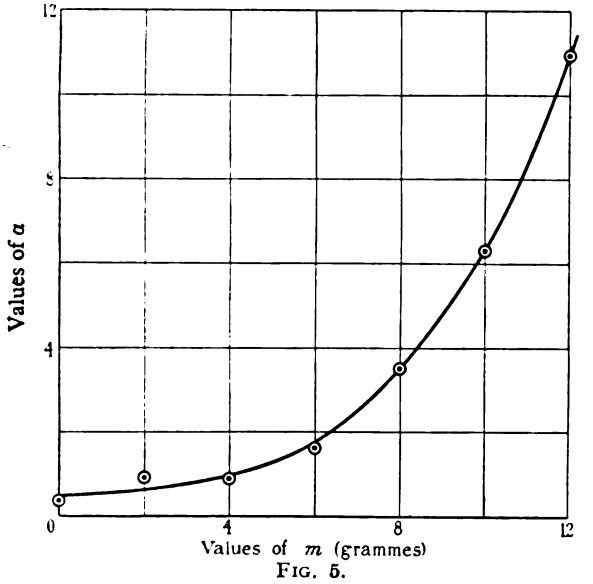

Fig. 5.

frequency, and was corrected as described above by the addition of a small mass to the prong carrying the free paddle.

The results, giving the total mechanical resistance of the two prongs (as determined from the resonance curve multiplied by 2) under different loads, are given below, the resonant frequency being constant at 103·2 cycles per sec.

	Resistance, in dynes/cm/sec.
(1) Thick oil round left-hand paddle with right-hand paddle free	2 776
(2) Thick oil round right-hand paddle with left-hand paddle free	3 320
Sum	6 096
(3) Thick oil round both paddles	5 980
(4) Both paddles free	136
Sum	6 116

Assuming no increase of prong resistance with unbalanced resistive load, the two sums should agree as they are each equal to the mechanical resistance of 4 prongs plus 2 immersed paddles plus 2 free paddles, and from the results it is clear that the assumption is justified.

From the balancing mass required with only one paddle immersed, a figure may be obtained expressing the " mechanical power factor " of the damping. With the right-hand paddle immersed, 1·0 gramme had to be added to the left-hand paddle to restore the mass balance.

The mechanical resistance of the oil damping with the right-hand paddle

$$= 3\,320 - 136 = 3\,184 \text{ dynes/cm/sec.}$$

Hence " power factor " of the oil damping

$$= \frac{r'}{\sqrt{[(r')^2 + \omega_0^2 (m')^2]}} = \frac{3\,184}{\sqrt{[3\,184^2 + (2\pi \times 103{\cdot}2 \times 1)^2]}} = 0{\cdot}98$$

This result shows that oil damping, at any rate under the conditions used, acts practically as a pure resistance.

APPENDIX 3.

Determination of the Force Factor A.

The force factor was determined in the first instance in the well-known manner by drawing curves of vibration amplitude against frequency and of coil impedance against frequency. The decay factor α is determined from either; the former gives, in addition, the amplitude x_0 at resonance and the latter the diameter OD in ohms of the motional impedance circle.

From (7), $$\omega_0 x_0 = \frac{A i_m}{r} \quad . \quad . \quad . \quad . \quad . \quad (9)$$

and $$\text{OD} = \frac{2A^2}{r} \quad . \quad . \quad . \quad . \quad . \quad (10)$$

whence $$A = \frac{\text{OD}}{2\omega_0 x_0} i_m \quad . \quad . \quad . \quad . \quad . \quad (11)$$

Subsequently, however, the determination of m and s from a curve such as Fig. 3, and α from a resonance curve of amplitude against frequency, enabled r to be found as

$$r = 2m\alpha$$

and a quicker determination of A from equation (9) as

$$A = \frac{\omega_0 x_0 r}{i_m} \quad . \quad . \quad . \quad . \quad . \quad (12)$$

For the experiments described in Section (3),

$\omega_0 = 2\pi \times 119$ radians per sec.
$x_0 = 0{\cdot}0108$ cm
$r = 15{\cdot}6$ dynes per cm per sec.
$i = 0{\cdot}176$ mA (R.M.S. value)

giving $A = 5{\cdot}06 \times 10^6$ dynes per abamp.

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APPENDIX 4.

Check on Constancy of Force Factor with Unbalanced Resistive Loads.

Using the equation (12) derived above, the force factor A was determined for the unbalanced resistive loads used in Appendix 3, giving the results tabulated below.

From the figures in the last column of the table it is clear that A is sensibly independent of the load on the fork, provided that there is mass balance.

	ω_0	x_0	r_1	i(R.M.S.)	A
	radians/sec.	cm	dynes/cm/sec.	mA	dynes/abamp
(1) Thick oil round left-hand paddle with right-hand paddle free	650	0·00794	1 388	10·1	$5{\cdot}00 \times 10^6$
(2) Thick oil round right-hand paddle with left-hand paddle free	650	0·0081	1 660	12·5	$4{\cdot}94 \times 10^6$
(3) Thick oil round both paddles	650	0·00685	2 990	19	$4{\cdot}91 \times 10^6$
(4) Both prongs free (from Appendix 3) ..	—	—	—	—	$5{\cdot}06 \times 10^6$

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SOME NOTES ON THE DESIGN OF A GRAMOPHONE PICK-UP.*

By G. W. SUTTON, B.Sc.

(*Paper first received 29th November, and in final form 20th December, 1929; read before the* WIRELESS SECTION *5th February, 1930.*)

SUMMARY.

It is a matter of common experience that the performance of the large majority of commercial pick-ups leaves much to be desired. The following notes describe an experimental investigation of the performance of certain types.

Further experiments are then described on the mode of vibration of a pick-up armature and on other factors affecting the performance of the instrument.

A theoretical treatment based on the results of these experiments is given and finally a new design is evolved.

TABLE OF CONTENTS.

(1) Classification of types of pick-up.
(2) Results of preliminary tests.
(3) Experiments on mode of vibration of a pick-up armature.
(4) Parasitic vibrations in the pick-up support.
(5) Lamination of armature and pole-shoes.
(6) Distribution of alternating component of the magnetic field at different frequencies.
(7) Effect of annealing armature and pole-shoes.
(8) Damping the vibrating system.
(9) Acoustical damping.
(10) Résumé of theory of operation.
(11) Method of design.
(12) Regulation of voltage output.

Appendix 1. Measurement of voltage output over a wide range of frequency.
Appendix 2. Theory of the pick-up.
Appendix 3. Tendency of the pick-up needle to leave the record groove.

(1) CLASSIFICATION OF TYPES OF PICK-UP.

It would appear that there is a demand for three types of pick-up:—

(1) A type, frequently called a "play-back," which is used to reproduce directly from the "wax" immediately after an impression has been made, in order to ascertain whether the recording is satisfactory. In this device voltage output is of relatively little importance compared with the necessity for an absolute minimum of wear by the needle.

(2) An instrument which is intended to be used with its own special amplifier, where again the voltage output is of minor importance since it can be so readily augmented in the amplifier.

(3) A type which can be used with a standard broadcast receiving set. This should provide an output of the order of 0·15 volt/mm/cycle/sec.† in order to comply with the present tendency in receiver design.

It is with type (3) that the present paper is primarily concerned.

It may be accepted at the outset that the technique of recording has advanced to a very high standard. The frequency range covered is such as to call for reproducing apparatus capable of dealing properly with frequencies down at least to 50 cycles per sec., and up to the region of 8 000 cycles per sec. In the author's opinion the importance of the lower frequency limit is still not fully appreciated.

Not only has the useful frequency band been very considerably widened, but the amplitude of the groove has been increased, at least in the better-class records. This means that the ratio of recorded note amplitude to that of those unavoidable irregularities which give rise to surface noise or "scratch" has been increased. Full advantage of this can only be taken by the use of electrical pick-ups, since some control of volume of the output becomes necessary. In fact there have been many protests from musically inclined listeners who use mechanical sound-boxes, at the excessive volume given by good modern records.

Considerations of voltage output and of cost of production led the author to conclude that the moving-iron electromagnetic type of generator was the only one likely to prove satisfactory in class (3). In designing this type we have to choose between the methods of construction which are generally known as the "attracted reed" and the "balanced armature." Of these the latter appears to possess two distinct advantages. Firstly, the resultant forces of magnetic attraction on the armature are smaller, and therefore do not necessitate the use of such a high order of stiffness in the armature suspension. Secondly, a comparatively much larger amplitude of armature vibration can be allowed without introducing "rectification" effects in the voltage output. These two points allow a very useful margin of freedom in deciding the dimensions of the armature and pole-pieces and of the constants of the principal vibrating system.

(2) RESULTS OF PRELIMINARY TESTS.

The results of some preliminary tests, using the method outlined in Appendix 1, on various commercial pick-ups, brought out the following points:—

(1) The existence, in every case, of one or more "resonant" points in the voltage characteristics. The prominence of these peaks varied considerably, and in some cases so affected the output that when the instrument was used for the reproduction of speech and music a clearly noticeable distortion resulted. It appeared probable that corresponding damage was

* Reprinted from *Journal I.E.E.*, 1930, vol. 68, p. 566.

† Measured from + peak to − peak of the amplitude of needle-point displacement.

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caused to the record. Such a case is illustrated in Fig. 1.

(2) A tendency, with some pick-ups, for the needle to jump out of the groove at frequencies below about 100 cycles per sec. In all cases a distinct vibration of

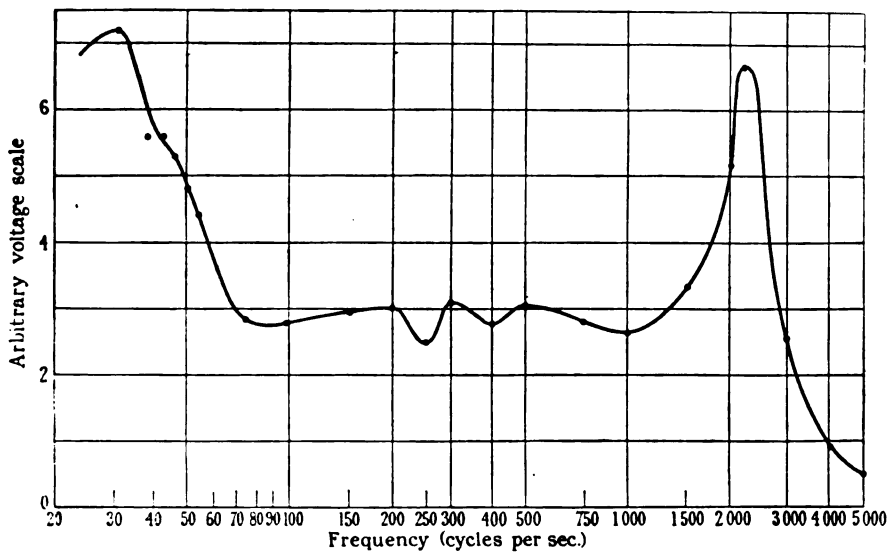

FIG. 1.—Voltage output characteristic of commercial pick-up using "loud" tone H.M.V. needle, showing two principal resonances.

the tone-arm or supporting lever could be felt at low frequencies.

(3) A comparatively poor voltage output at frequencies of 5 000, 4 000 and sometimes even 3 000 cycles per sec. This leads to a distinct lack of "brilliance" in speech and music reproduction.

(3) EXPERIMENTS ON MODE OF VIBRATION OF A PICK-UP ARMATURE.

The reason for the peak in the characteristic in Fig. 1 at 2 200 cycles is not at first clear. It is true that at a certain frequency the reaction force due to the moment of inertia of the vibrating armature is balanced by that due to its method of suspension, leaving only frictional or damping forces at play. This, however, should only increase the amplitude of vibration at that frequency *for a constant deflecting force*. When the needle-point rests in the groove of a record the amplitude of excursion of the point is governed, however, by the record, and the deflecting force adjusts itself accordingly. Consequently, providing the armature is rigid and the needle-point does not leave the groove there is no reason why there should be a peak in the voltage characteristic at this frequency. To explain the existence of the peak, therefore, it is necessary to assume either that the armature is not rigid or that the needle itself bends very considerably. It was thought worth while to investigate this point experimentally, and the following simple test was devised. The commercial pick-up of Fig. 1 was mounted in an inverted position in a heavy clamp. A certain measure of flexibility (to simulate that existing when the instrument is fitted to a tone-arm) was provided by blocks of rubber (R in Fig. 2). A mirror, about 2 mm square, stuck to the armature as close to the air-gap as possible, reflected a clearly defined spot of light on to a scale at a distance of about 3 m. By this means vibrations of so small a total amplitude as 0·00005 radian could be observed, and 0·0001 radian and upwards could be fairly accurately measured. A constant-amplitude variable-frequency current was passed through the pick-up windings by placing it in the anode circuit of a power valve excited by a heterodyne oscillator. The valve was of the "pentode" class, having a differential resistance of 30 000 ohms. This ensured a sufficient degree of constancy of the current with frequency (for constant voltage input to the grid) up to about 3 000 cycles per sec. Curve A in Fig. 3 shows the result of a test carried out under these conditions. One clearly defined

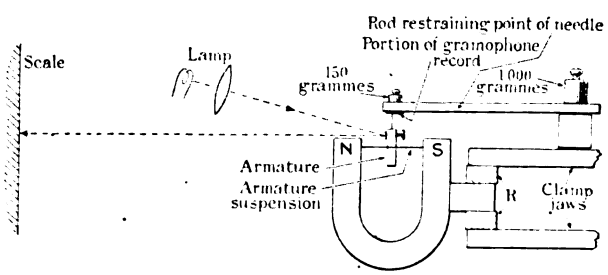

FIG. 2.—Method used for examining the mode of vibration of the armature.

resonance is noticed at 715 cycles per sec. The armature of this generator was of simple shape and its moment of inertia could readily be calculated. The value was 1·52 gramme-cm^2. The stiffness of the suspension was measured by hanging weights on a short length of steel wire clamped in the needle holder and observing the deflection. The mean value obtained was 23×10^6 dyne-cm per radian. From this and the moment of

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inertia, the resonant frequency is found to be 620 cycles per sec.; a result in fair agreement * with the value obtained from Fig. 3.

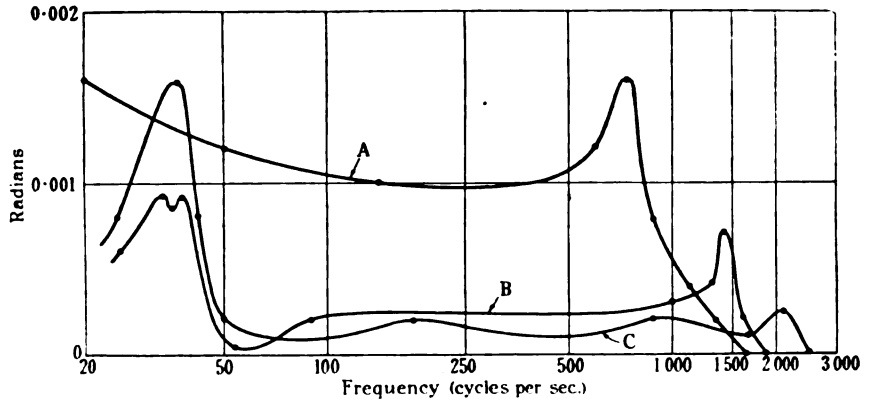

FIG. 3.—Amplitude of armature vibration of commercial pick-up. Vibrations produced electrically; constant input current.
A.—Armature free. B.—Needle-point restrained, "soft" needle. C.—Needle-point restrained, "loud" needle.

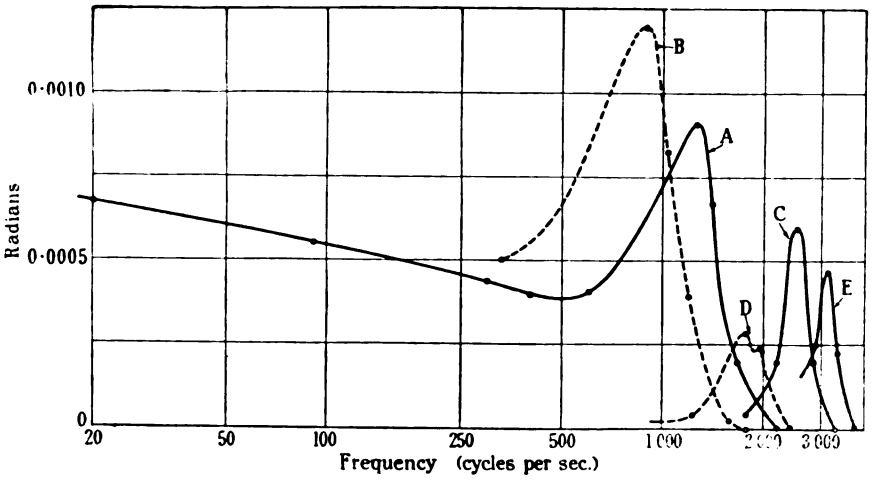

FIG. 4.—Amplitude of armature vibration of experimental pick-up
A.—Armature free. B.—Armature free, moment of inertia artificially increased. C.—Needle-point restrained, "soft" needle.
D.—Needle-point restrained, "soft" needle. E.—Needle-point restrained, "loud" needle.

It will be noted that there is no appreciable peak in the voltage characteristic (Fig. 1) at this frequency.

trated in Fig. 2. The 150-gramme weight corresponded to the weight of the pick-up when in its normal operating

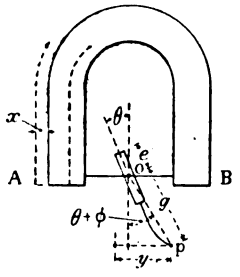

FIG. 5.

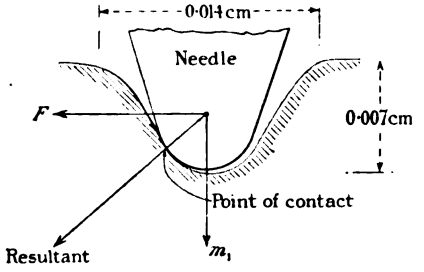

FIG. 6

The test illustrated in Fig. 2 was than extended by restraining the point of the needle in a manner which reproduced, as closely as possible, the conditions under

position. The readings obtained with the same operating current as before are shown in Fig. 3, curves B and C.

It will be seen that the general level of amplitude is

* Compare similar calculation on page 120.

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much reduced in curves B and C and that there is no sign of the peak at 620 cycles per sec. Instead, prominent peaks are introduced at 40 cycles and at 1 450 and 2 100 cycles for the "soft" and "loud" needles respectively. The two peaks in curve C correspond almost exactly to those in Fig. 1, and the test demonstrates fairly conclusively that the bending of vibrations transmitted by the pick-up along the lever would be absorbed instead of being reflected. The hole through the weight was therefore bored out to a larger diameter and the space so formed between it and the rod was packed with felt. This was sufficient to reduce the irregularity in the characteristic to the negligible proportions illustrated in curve C.

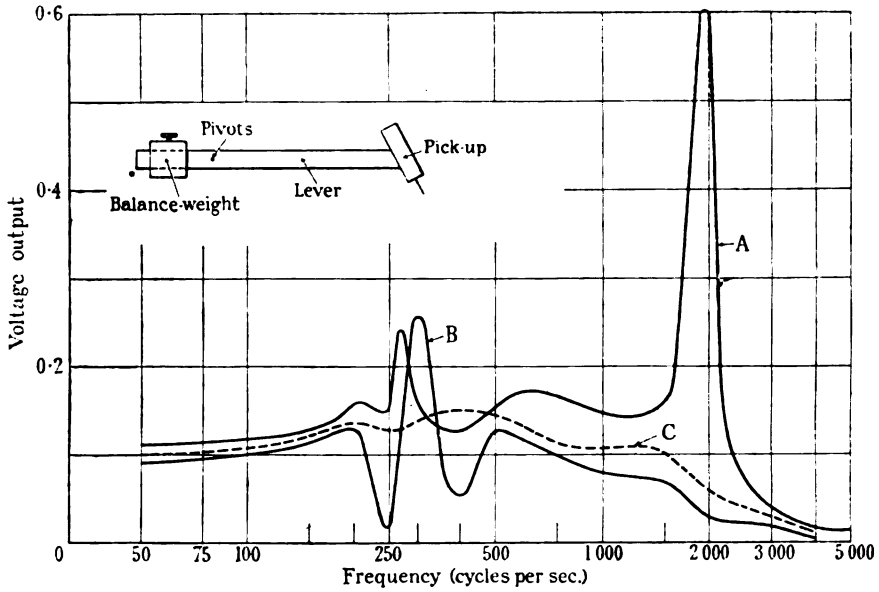

FIG. 7.—Effect of vibration in supporting lever. Balanced-armature pick-up; armature mounted on 0·0085 in. disc.

A.—No added damping. B.—Heavy oil damping. C.—Medium oil damping; balance-weight insulated from lever with felt.

the needle plays an important part in the performance of the amplifier, particularly at the higher frequencies.

(4) PARASITIC VIBRATIONS IN THE PICK-UP SUPPORT.

With a simple and rigid construction no complications will arise from parasitic vibrations in the armature itself, except at quite high frequencies, i.e. above 5 000 cycles per sec. On the other hand, trouble may readily arise at lower and more important frequencies from vibrations of the pick-up support or of some part of it. Evidence of this was apparent in several of the output characteristics taken in the course of the work, and the most striking example is illustrated in Fig. 7. In the characteristic A a peculiar irregularity occurs between 200 and 500 cycles per sec. The shape resembles that of the impedance/frequency curve of a coil to which a tuned circuit is coupled, in the neighbourhood of the resonant frequency of the latter. It suggests, therefore, that energy is being absorbed by some system coupled to the pick-up. It was noticed also that the counterbalance weight vibrated appreciably at about 300 cycles per sec. When damping oil was put into the pick-up the effect was very considerably intensified (see curve B).

The counterbalance weight was a sliding fit on the supporting lever and was held in position by a clamping screw. It was thought that by insulating this weight from the rod instead of clamping it rigidly to it the

(5) LAMINATION OF ARMATURE AND POLE-SHOES.

In all apparatus designed for alternating fields, particularly of high frequency, care is taken to laminate the iron core. It is a matter of some difficulty to do this in the present case owing to the limitations of size.

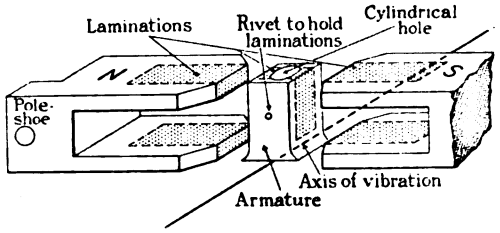

FIG. 8.

Experiments were, however, made to ascertain whether improved performance at 3 000, 4 000 and 5 000 cycles per sec. could be obtained by this means.

The shape of the armature and pole-shoes of the model upon which the experiment was conducted is shown in Fig. 8. The armature consisted in the first place of a short length of annealed mild-steel rod, hollowed out by drilling as shown. A voltage characteristic of this instrument was taken, after which it was disassembled and the portions shown dotted were cut away and replaced by thin stalloy laminations. After reassembling a fresh characteristic was taken. It was

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found that only slight improvements in voltage and output resulted and that this was more pronounced for frequencies *below* 200 cycles than for those above it. In any case it was not sufficient to justify the complications of construction, and it appeared that improvements at the higher frequencies were more likely to be obtained from reduction in the inertia of the armature. The apparent anomaly of a greater improvement at lower than at higher frequencies may indicate insufficient insulation between laminations; in the author's opinion, however, it is more likely to be due to the difficulty of reassembling such a small instrument to exactly the same gap dimensions and motional constants of the armature as it previously possessed.

(6) Distribution of Alternating Component of the Magnetic Field at Different Frequencies.

The experiment just mentioned suggested that the falling off at high frequencies was not due in such large measure as had previously been supposed to the effect of eddy currents on the alternating flux distribution. A further experiment to check this point was carried out.

A coil was wound in three sections (a, b and c, Fig. 9) of 2 000 turns each, and the voltage outputs from the three sections were measured and compared. As will be seen from the diagram, the changes in flux distribution due to the motion of the armature will be greatest at the pole-tips. Consequently a given amplitude of vibration will induce a larger voltage in "a" than in "c"; and the higher the frequency the greater will be the ratio of V_a to V_c. In fact it had previously occurred to the author to make use of this by backing-off V_c against V_a and V_b in series, with the object of levelling up the voltage characteristic of the pick-up. Measurement showed, however, that the difference was not actually as great as was anticipated. The figures in Table 1 illustrate the point.

Table 1.

Cycles per sec.	50	100	250	500	1 000	5 000
V_a volts	0·061	0·071	0·094	0·091	0·075	0·029
V_b volts	0·061	0·066	0·085	0·082	0·065	0·023
V_c volts	0·056	0·064	0·083	0·082	0·065	0·016

Table 2.

Cycles per sec.	50	150	500	3 000	4 000	5 000
Output before annealing, volts ..	0·140	0·137	0·164	1·50*	0·101	0·055
Output after annealing	0·167	0·207	0·220	3·0*	0·155	0·084
Percentage increase	19	51	34	100	55	53

* No additional damping was introduced for this experiment, and 3 000 cycles was the principal resonant frequency.

(7) Effect of Annealing Armature and Pole-Shoes.

The opportunity was taken when constructing one model to ascertain the extent of the improvement in output which might be expected from annealing the mild-steel portions of the magnetic circuit. The difficulty in making the experiment was to reconstruct the instrument after annealing to the same dimensions as it possessed originally. In order to reduce as much as possible the uncertainty due to this, the model selected was one which had been constructed with comparatively long air-gaps.

The armature was a short length of mild-steel rod and the pole-shoes were machined from a sheet of the same material. The annealing process consisted of heating the portions in a gas flame to a pale yellow colour and then allowing them to cool slowly by gradually reducing the flame. The operation was spread over a period of about ½ hour.

The results are given in Table 2.

(8) Damping the Vibrating System.

When the reaction force due to the inertia of the armature is balanced by that due to the stiffness of the needle and the armature suspension, the damping alone controls the extent of the vibration. As the frictional

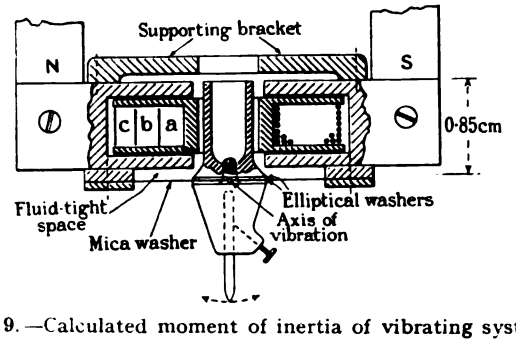

Fig. 9.—Calculated moment of inertia of vibrating system:

Armature (above axis)	0·08	(mass 0·55 gramme)
Needle-clamp	0·08	(mass 0·87 gramme)
Clamping screw	0·17	(mass 0·35 gramme)
Needle	0·12	(mass 0·19 gramme)
	0·45	gramme-cm

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forces are ordinarily much smaller than the reaction forces, both voltage output and record wear become excessive at this resonant frequency. Some method of increasing the damping is therefore essential. The means available may be classed as follows:—

(1) To raise the resonant frequency to the neighbourhood of 4 000 or 5 000 cycles.

(2) To couple to the armature a second vibrating system which absorbs energy, either electrically or mechanically.

(3) To increase artificially the damping factor by mechanical means.

(1) If this can be achieved by a reduction of the moment of inertia of the armature it is the most satisfactory way of dealing with the problem. Both the damping "constant" itself, for a given construction, and the force corresponding to a given damping constant, increase with frequency. Consequently, if the resonant frequency is raised to, say, 5 000 cycles per sec., additional damping becomes unnecessary, that inherent in the suspension being sufficient. A sufficient reduction in the inertia of the armature can only be achieved, however, at such a sacrifice of voltage output as to remove the pick-up from the present class as regards sensitivity.

It is much more common to attempt to get the same result by inserting small pieces of rubber between the armature and pole-faces. This is usually spoken of as "damping" the system with rubber. Actually, however, the stiffness of the system is increased to a far greater extent than the damping factor, and this has an undesirable effect on the performance of the pick-up at low frequencies (see equations 8 and 11).

(2) Unfortunately, the degree of coupling between the mechanical vibrating system of the pick-up and the electrical circuit is not sufficiently close to make it feasible to damp the former with a tuned-circuit shunt. It is necessary, therefore, to employ a second mechanical system and to couple it mechanically with the first. A form of construction which makes this possible is described later. Preliminary experiments with the method have given promising results.

(3) It is most commonly attempted to employ rubber for this purpose, as mentioned above. The difficulty of devising a satisfactory construction to this end is considerable. In the case of an electromechanical recorder* it was found necessary to use a rod of gum rubber 25 cm long to provide a non-reactive load sufficient to swamp the reactive forces set up in the vibrating system during operation. The rod was attached to the armature in such a manner that torsional vibrations were set up in it. These were transmitted axially from end to end, and were absorbed by the time they had travelled along the rod and back again. It is apparent, therefore, that if vulcanized rubber, which is presumably far less viscous in nature than gum rubber, is to be employed, its dimensions and disposal must be exceedingly carefully worked out.

The use of viscous liquid, such as a heavy oil, offers a simpler solution to the problem. The difficulty which immediately arises, however, is to design a suitable liquid-tight container.

* *Bell Telephone Laboratories Reprint*, B—377 ,3, p. 5.

At least one good commercial pick-up has been constructed on these lines and one of the designs described later employs the same method.

The author has also used the friction set up between two small oily mica surfaces, one carried by the armature and one attached to the pole-shoes. This has proved satisfactory, but suffers from the drawback of needing very careful and frequent adjustment.

"Book" damping has also been tried. This consists of a number of small sheets of tissue paper which are alternatively compressed and released by the motion of the armature. In the author's experience this method also needs frequent adjustment owing to the liability of the paper to become "packed."

(9) Damping Due to Acoustical Output from a Mechanical Sound-Box.

It is well known that the performance, as judged by ear, of a modern mechanical sound-box connected to a well designed tone-arm and horn is remarkably good, despite the somewhat complicated nature of the lever system by which the needle operates the diaphragm, and the fact that the only damping incorporated is that due to the acoustic load. It was thought worth while, therefore, to study the vibration of the operating lever of a sound-box with and without its acoustic load.

The method employed was to mount the permanent magnet, pole-tips and coils of a telephone ear-piece on the outer face of the sound-box. A very small piece of annealed mild steel was firmly attached to the sound-box lever close to the centre of the diaphragm, and the ear-piece was clamped in such a position that the piece of steel formed its armature. Thus the vibration of the lever produced proportional voltages in the ear-piece coils, and these were measured on the valve voltmeter as before. The instrument became, in fact, a combined sound-box and pick-up. As such it was useful for the comparison by ear or microphone of the acoustical output from the gramophone itself with that from the amplifier and moving-coil loud-speaker. By means of a switch in the pick-up circuit, the loud-speaker reproduction from a record could be superimposed on the direct gramophone reproduction, or removed, at will.

In Fig. 10, curve A shows the voltage generated by the sound-box pick-up when operating under normal conditions. Over the frequency range employed it will represent fairly closely the amplitude of vibration of the centre of the sound-box diaphragm. Curve B was taken after the narrow end of the tone-arm had been tightly plugged with cotton-wool, thus preventing sound output through the horn. A prominent resonance point is apparent, under these conditions, at about 1 300 cycles per sec. The effect of the acoustical load in obliterating this resonance is apparent from comparison of the curves. If impedance/frequency characteristics of a horn-type moving-iron loud-speaker are plotted with and without the horn in place, it is seen that the resonance effect of the armature is smothered by the acoustical damping in just the same way.

The curves of Fig. 10 also indicate how closely the mechanical and acoustical systems of a well-designed gramophone are coupled, since the numerous horn resonance points impress themselves upon the vibrations

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of the lever system. When the same tone-arm and horn were used in conjunction with a different sound-box, the cavities in which were not so well adjusted to the aperture in the tone-arm, this effect was not nearly so pronounced, although a useful degree of damping was provided by the acoustical load. The results of a measurement under these conditions are given in Fig. 11.

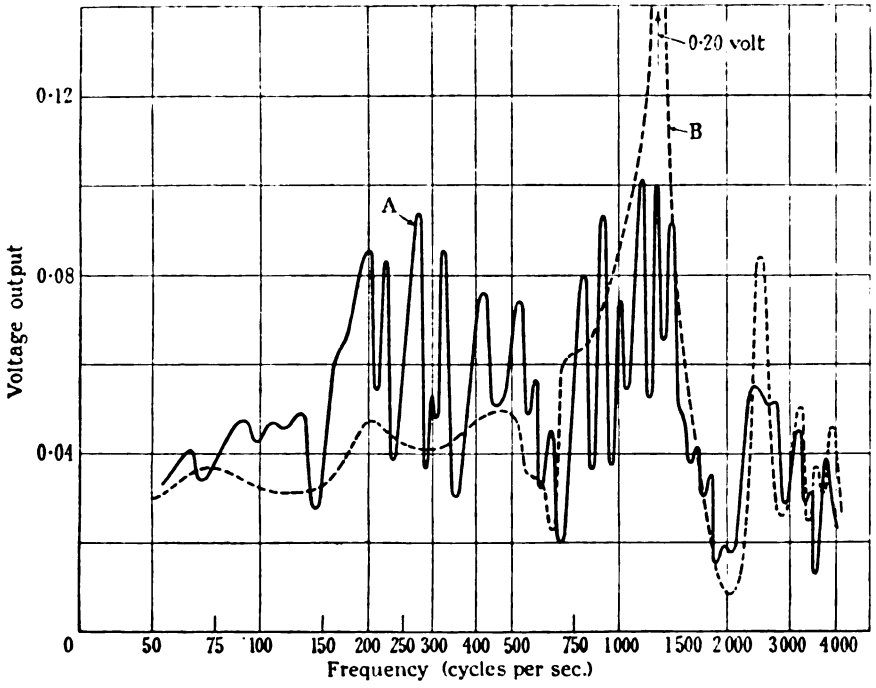

FIG. 10.—Sound-box lever as armature of electromagnetic generator.

A.—Sound-box, tone-arm and horn in normal operation. B.—Acoustic output prevented by plugging tone-arm with cotton-wool.

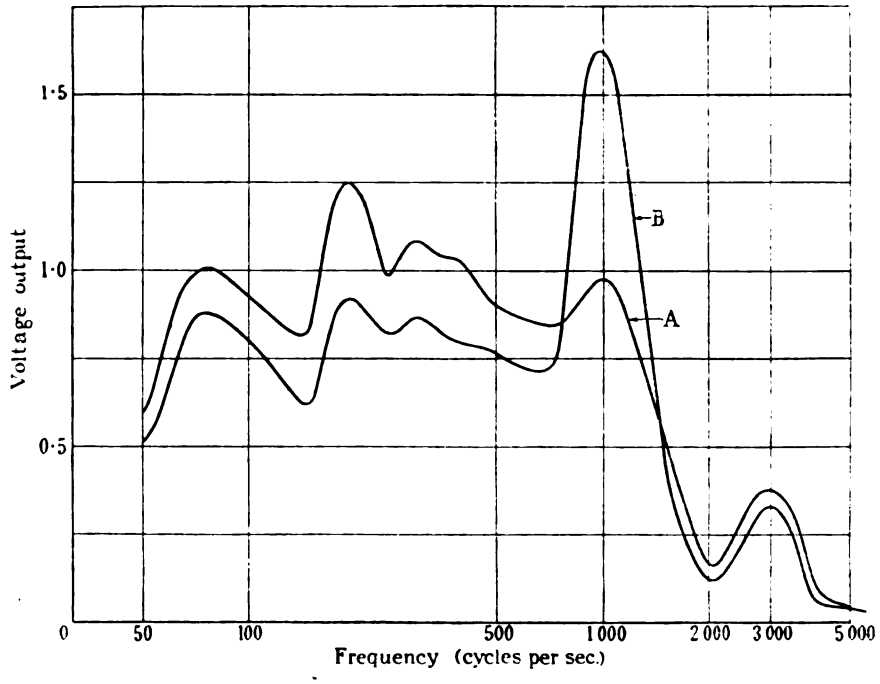

FIG. 11.—Voltage output of sound-box pick-up.

A.—When loaded acoustically. B.—With tone-arm closed, preventing the full acoustic output.

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In comparing the curves of Fig. 10 with those of Fig. 11 it is noticeable that the output falls off above 2 000 cycles much more in the latter case than in the former. This was doubtless due to the greater rigidity of the lever system in the former. An appreciable lack of brilliance in the reproduction from the second sound-box, both in the case of the electrical and the mechanical methods, was very apparent.

These curves show clearly, when compared with Fig. 12, the advantage to be gained by the use of the simple rigid vibrating system of the electrical pick-up as compared with the more complicated one necessary in the case of the sound-box.

(10) Résumé of the Theory of Operation.

In Appendix 2 the equations of motion of the vibrating armature of a pick-up are examined and it is found that the position on the scale of frequency of the principal resonance is governed by the inertia of the armature and the sum of the stiffnesses of the armature suspension and of the needle. In a well-designed instrument the stiffness of the suspension will be small compared with that of the needle (of the order of one-tenth), so that the latter is virtually the governing factor. In fact the only means available to the designer to raise the principal resonance to a high figure (as suggested in Section 8) is to reduce the armature inertia.

It is also seen that a state of resonance arises at a very low frequency, the pick-up vibrating as a whole about the point of the needle and the pivot of the tone-arm or lever. This frequency is governed by the stiffness of the suspension and the inertia of the body of the pick-up about the axis of suspension of the armature. It is not a difficult matter, therefore, to reduce it to so low a figure that its effect is of no importance, by increasing the mass of the pick-up and its support. Except for the two resonant conditions just mentioned, there appears to be no reason why the amplitude/frequency characteristic of the armature of a pick-up should not be a horizontal straight line up to the point at which it is no longer justifiable to assume it to be a rigid structure.

An attempt is made in Appendix 3 to deduce the conditions under which the needle is liable to leave the groove.

It is seen from equation (11) that the inertia of the pick-up about the support pivots should be large, so that $1/(\omega^2 m)$ may be small compared with $g^2/\mathbf{T}$. Also that $\mathbf{T}$, the suspension control constant, should be as small as possible so that $m_1 g$, the bearing pressure on the needle, may be reduced to a minimum.

It is a common experience to find that in a commercial pick-up which has been " damped " by inserting small rubber blocks in the air-gap, the value of $\mathbf{T}$ has been raised to such a value that the needle will not follow the groove of a test record at 100 or even 150 cycles per sec. without holding it down (i.e. raising $m_1 g$ considerably).

(11) Method of Design.

The reasons for adopting the balanced armature system have already been given in Section (1). It remains to deduce from the foregoing data and those in Appendices 2 and 3 the best dimensions and constants for the vibrating system.

The most important single property of this system is its moment of inertia, and every effort must be made to reduce it to a minimum. The system must consist of the needle itself, needle-clamp and armature. In the case of types (1) and (2) (see page 112) it is possible to dispense with a separate portion for the armature and to use either the needle or the clamp for the purpose. An attempt was made to use the clamp in this double role in the first model which the author constructed to fulfil the conditions of type (3), but it was then found difficult to incorporate a damping device within the limited space involved, and the attempt was abandoned.

It is an advantage to suspend the system at its mid-point, since for a given mass this leads to the lowest moment of inertia and also facilitates the application of additional damping.

Analysis of the performance of the commercial instrument of Fig. 1 leads to the conclusion that its inertia constant is much too high (1·5 gramme-cm^2). The experimental armature was therefore designed so that $\mathbf{I}$ should be not more than 0·5 gramme-cm^2. This was achieved by making the armature portion hollow (see Fig. 9). It is well known that a given mass distributed in the form of a hollow cylinder or rectangular block will possess much greater rigidity than the same mass in the form of a rod. Two further advantages accrue from this construction. Skin effect would in any case

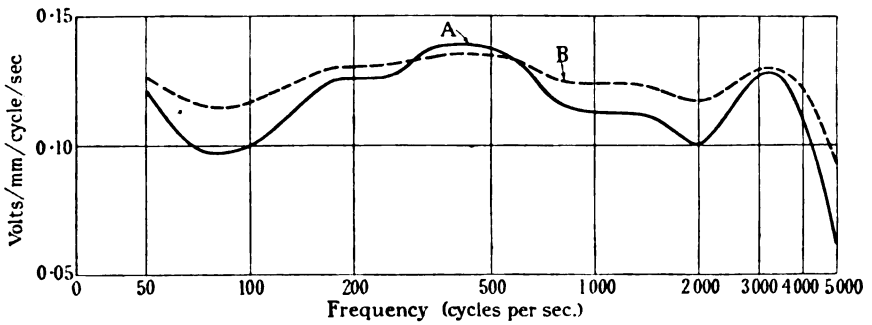

Fig. 12.—Voltage characteristic of experimental model used with a 6 : 1 step-up transformer.

A.—To a scale of volts (measured at secondary of transformer). B.—To a scale of log (volts).

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restrict the flux to the surface of the armature at the higher frequencies. Hollowing it out will therefore tend to produce a more uniform characteristic; while the greater diameter of armature involved lengthens the magnetic paths straight from pole-face to pole-face for a given length of air-gap between armature and pole-face.

The hollow in the armature has also been used for coupling the system with rubber or oil (i.e. " capacity " or " resistance " coupling) to a second vibrating system mounted above the first, to provide increased damping.

The armatures in the experimental models subsequently built were mounted on mica discs as shown in Fig. 9. Mica was used since it is non-magnetic and has a high stiffness/mass ratio. Moreover, having decided on a suitable thickness (0·010 in. was employed finally) the discs are easily made to specification and are very uniform in their properties.

In order to prevent motion of the armature in a plane parallel to the pole-faces due to irregularities in the record groove, two elliptical washers were clamped into the positions shown in Fig. 9. The minor axes of these were made equal to the armature diameter, and the major axes were made about twice as long. Measurement showed that this rendered the stiffness of the support in the plane parallel to the pole-faces nearly twice that in the plane of normal armature vibration.

The mica disc was clamped at its outer edge between two heavy rings, with carefully turned flat faces, and these in their turn were bolted to the pole-pieces (Fig. 9).

The thickness of the disc was adjusted by experiment until it was just sufficient to prevent the armature sticking to the poles by magnetic attraction. The value of **T** was then found to be $13{\cdot}8 \times 10^6$ dyne-cm per radian, which is little more than half the value for the commercial instrument previously tested. The above value was determined by deflecting the armature by a known torque in the usual manner. It appears probable that at high frequencies the effective value of **T** increases due to the mode of vibration of the disc and the resilience of the air in the cavities around the armature, since the values obtained from the tests illustrated in Fig. 4 are considerably higher than this. Thus the resonant frequency shown in curve A would be 890 cycles per sec. from $\mathbf{T} = 13{\cdot}8 \times 10^6$ dyne-cm per sec. and $\mathbf{I} = 0{\cdot}45$ gramme-cm^2, whereas it is actually 1 250 cycles per sec., indicating either an increase in the effective value of **T** at these frequencies, or an error in calculating **I**.

The values of **I** calculated from these curves, in terms of a known increment of moment of inertia (attaching a known weight at a measured distance from the axis), were 0·42 and 0·51 gramme-cm^2 (0·45 gramme-cm^2 calculated from dimensions). It was difficult to measure accurately the effective radius of gyration of the added mass, owing to the small dimensions of the armature and the irregularity of its contour. The value of **T** calculated from the curves, using $\mathbf{I} = 0{\cdot}45$ gramme-cm^2, is 30×10^6.

It has been seen (equation 11) that **T** should have as small a value as possible to reduce the tendency for the needle to leave the groove and the consequent wear and tear on the record, at frequencies of the order of 100 cycles per sec. At these low frequencies it is safe to assume that the effective value of **T** approximates much more closely to the value $13{\cdot}8 \times 10^6$ (obtained statically) than 30×10^6 (obtained dynamically at 1 000 cycles per sec. and over). This is confirmed by the comparative freedom of this model from tone-arm vibration at 100 cycles per sec.

That the value of **T** (measured statically) was sufficient to prevent amplitude distortion is indicated by the linear relationship found to exist between deflecting torque and angular displacement for the armature, up to values far in excess of the amplitudes obtained in operation, even at low frequencies.

The air-gap length was about 0·3 mm, whereas the maximum amplitude of displacement of the armature in the gap was about 0·05 mm.

Having thus determined **I** and **T**, the remaining important quantities are the mass of the pick-up itself and the bearing pressure on the needle. From consideration of equation (11) it is seen that the former should be as large as possible *—of the order of 300 or 400 grammes—while the latter should be no more than is sufficient to prevent the needle leaving the groove at the lowest frequencies and the largest amplitudes of vibration. These conditions may be fulfilled by connecting the pick-up very rigidly to a massive and rigid supporting lever. The weight of the lever should be so disposed about its pivots that the bearing pressure on the needle is about 100 grammes. This, assuming the given values of g and **T**, provides a sufficient margin of safety.

It will be seen from Fig. 9 that the construction of this model leaves a space above the mica disc and around the armature which may be rendered liquid-tight by varnishing, and may then be filled with oil of suitable viscosity to provide the necessary additional damping.

The voltage characteristic for this instrument is given in Fig. 12, both to a scale of volts/mm/cycle/sec. and also to the logarithm of this quantity.

(12) Regulation of Voltage Output.

It is usual to employ a simple resistance potential-divider for this purpose. Unless its ohmic resistance is very high, however, this will cause considerable falling off of output in the upper part of the frequency scale.

The coil of the experimental instrument just described had 6 000 turns and an inductance, measured at 500 cycles per sec., of 2·25 H. Neglecting its resistance, this corresponds to an impedance of about 700 ohms at 50 cycles and 70 000 ohms at 5 000 cycles. Consequently even a 70 000-ohm potential divider would lead to a 50 per cent loss of output voltage at 5 000 as compared with 50 cycles per sec. A tapped choke, having a liberal iron core in order to prevent excessive voltage-drop at low frequencies, and an inductance of at least 10 H, would be preferable for the purpose.

It is even better to use a transformer with a tapped secondary. The pick-up can then be wound with a few hundred turns of heavier-gauge wire, thus reducing the tendency to oscillation of the amplifier when long leads are used to connect the two.

* Thus making the term $1/(\omega^2 m)$ in equation (11) negligible compared with $g^2/\mathbf{T}$ at 100 cycles per sec.

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ACKNOWLEDGMENT.

The work was carried out at the Telegraphy and Telephony Laboratories of the City and Guilds (Engineering) College, and the author is indebted to Prof. C. L. Fortescue, O.B.E., Member, and to Prof. E. Mallett, Member, for facilities placed at his disposal and for helpful suggestions made.

APPENDIX 1.

PRELIMINARY EXPERIMENTS. THE MEASUREMENT OF VOLTAGE OUTPUT.

A set of sine-wave " constant power " records covering a range of frequencies from 50 to 5 000 cycles per sec. was employed to drive the pick-up. The amplifier and valve voltmeter for measuring the output presented little difficulty. The voltmeter was of a type previously described by the author.* The rapidity of indication of this type enables readings to be obtained in very rapid succession and with a minimum of wear on the record—an important point when a large number of comparative tests are carried out.

To obtain readings intermediate between the fixed frequencies (50, 75, 100 and 150 cycles per sec. and so on) provided by the records, the speed of the mechanical motor driving the record turn-table was varied by means of the mechanical governor. A range of speed of from 46 to 88 r.p.m. was thus rendered available. The speed at any given setting of the control lever was found to be remarkably constant. This is illustrated by the readings set out in Table 1.

The range of speed variation was of the greatest use in taking the characteristics, as it enabled points of inflexion in the curve to be accurately ascertained. For speeds other than that at which the record is intended to be run, correction of the voltage reading has, of course, to be made.

Although the product of amplitude and frequency was supposed to be constant over the whole of the records, it was soon suspected that this was only approximately the case. To ascertain the extent to which the product varied from one frequency to the next, the voltage output from a good pick-up on the 50-cycle groove when running at, say, 90 r.p.m. was compared with that from the 75-cycle groove running at 60 r.p.m. The frequency is the same in both cases and, provided the amplitude characteristic of the pick-up used is reasonably good, the ratio of the voltages is the ratio of the groove amplitudes. By a succession of such measurements each groove may be compared with the next in the series. The results of this test are shown in Table 2.

The figures given express the actual amplitude as a decimal of that which would render the product with frequency equal to the corresponding product for the 50-cycle groove.

The test was found to be difficult to carry out with accuracy as the groove for any one frequency varied in amplitude slightly from point to point, and accidental irregularities were troublesome. Moreover, the errors are obviously cumulative in ascending the frequency scale, and no simple method of check from, say, the 500-cycle to the 50-cycle groove suggests itself. The departures from the ideal in the matter of amplitude are so serious, however, that even an approximate correction is valuable.

The measurement of absolute amplitude of the grooves was carried out by driving the pick-up from a tuning fork which was excited by a valve oscillator. The vibration of the tip of the fork was observed and measured by a microscope whilst the needle of the generator was allowed to rest on a small piece of broken gramophone record firmly attached to the prong close to the tip. The amplitude of vibration of the fork was adjusted by controlling the current from the maintaining valve until the voltage generated by the pick-up was the same as that produced when it was driven by a record of the same frequency.

TABLE 1.

Lever setting	0	1	2	3	4	5	6	7
Speed (r.p.m.)	46·5	58·8	67·4	74·0	79·4	84·3	88·3	92·3
Speed (r.p.m.) taken at a different time	47·4	58·5	67·4	73·9	79·6	84·2	88·2	92·0

TABLE 2.

Frequency (cycles per sec.)	50	75	100	150	200	250	300	400	500	750	1 000	1 500	2 000	3 000	4 000	5 000
	1·0	0·91	0·78	0·80	0·47	0·42	0·66	0·50	0·40	0·53	0·53	0·56	0·67	0·56	0·48	0·32

APPENDIX 2.

THEORY OF THE PICK-UP.

We may consider the generator to have a total mass of m grammes, and to be mounted on a tone-arm which exerts negligible control and frictional forces. The armature is assumed to have a moment of inertia of **I** gramme-cm^2, and its suspension to exert a control torque of **T** dyne-cm per radian, together with a frictional force A dyne-cm per radian per sec. In addition,

* *Proceedings of the Physical Society of London*, 1927-8, vol. 40, p. 14.

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the armature itself is assumed to be rigid, whilst the needle is not rigid, i.e. it is capable of bending when its point is deflected. The effect of the stiffness of the needle on the armature may be represented by an additional control torque $\mathbf{T}_n$ dyne-cm per radian, the angle considered being that subtended by the point of the needle when bent, and its mean position, at the axis of vibration.

Let y be the instantaneous displacement of the point p (Fig. 5) by the action of the record groove. This will result in a bodily displacement x of the generator, an angular displacement θ of the armature, and an angular displacement $(\theta + \phi)$ of p about the axis 0.

Let the horizontal component of the forces exerted by the groove at this instant be F dynes. This is equivalent to an equal force acting horizontally through 0 and a couple Fg acting on the armature.

The reaction couple acting on the armature

$$= \mathbf{T}\theta + Ad\theta/dt + \mathbf{I}d^2\theta/dt^2 \quad . \quad . \quad . \quad (1)$$
$$= Fg$$

Also, along the axis AB we have

$$F = md^2x/dt^2 \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

and, for the bending of the needle,

$$Fg = \mathbf{T}_n\phi \quad . \quad . \quad . \quad . \quad . \quad . \quad (3)$$

If the displacement of p varies harmonically and the resultant reaction forces are as given in the above equations, the force exerted by the groove will be harmonic and so also will the displacement x, θ and ϕ.

Therefore we may write

$$F'g = \mathbf{T}\theta' + jA\omega\theta' - \omega^2\mathbf{I}\theta' \quad . \quad . \quad . \quad (4)$$
$$= a\theta'$$

and

$$F' = -\omega^2 mx' = -\beta x' \quad . \quad . \quad . \quad (5)$$

Where F' and θ' are vectors representing harmonically varying quantities.

The alteration in gap length due to the above displacement is $= e\theta$, and the movement of the needle-point is

$$x + g(\theta + \phi)$$

$e\theta/[x + g(\theta + \phi)]$ may be called the "output factor" of the pick-up, and would have a value e/g, independent of frequency, in the ideal instrument.

Taking R.M.S. values we have

$$\frac{e\theta'}{x' + g(\theta' + \phi')} = \frac{e(g/a)F'}{F'/\beta + g(F'g/a + F'g/\mathbf{T}_n)}$$
$$= \frac{eg/a}{1/\beta + g^2(1/a + 1/\mathbf{T}_n)}$$
$$= \frac{eg}{\{(g^2/\mathbf{T}_n - 1/\omega^2 m)\}(\mathbf{T} - \omega^2\mathbf{I} + jA\omega) + g^2}$$

Simplifying by putting $e = g$ and re-arranging, we have

$$\frac{e\theta'}{x' + g(\theta' + \phi')} = \frac{\mathbf{T}_n\omega^2 mg^2}{[(\omega^2 mg^2 - \mathbf{T}_n)(\mathbf{T} - \omega^2\mathbf{I}) + \mathbf{T}_n\omega^2 mg^2] + jA\omega(\omega^2 mg^2 - \mathbf{T}_n)} \quad . \quad . \quad . \quad . \quad (6)$$
$$= \frac{\mathbf{T}_n\omega^2 mg^2}{y + j\delta}$$

the amplitude of which is

$$\frac{\mathbf{T}_n\omega^2 mg^2}{\sqrt{(y^2 + \delta^2)}}$$

The values of the quantities involved, in the case of the generator of Fig. 1, are as follows:—

$m = 135$ grammes, $\mathbf{I} = 1{\cdot}52$ gramme-cm^2, $g = 2$ cm,
$\mathbf{T} = 23 \times 10^6$ dyne-cm per radian

At 3 000 cycles per sec, $\omega^2\mathbf{I} = 5{\cdot}4 \times 10^8$ and $\omega^2 mg^2 = 19{\cdot}2 \times 10^{10}$.

It is assumed, and is confirmed later, that at this and neighbouring frequencies $\mathbf{T}_n$ is negligible in comparison with $\omega^2 mg^2$, and y (equation 6) then reduces to

$$\omega^2 mg^2(\mathbf{T} - \omega^2\mathbf{I}) + \mathbf{T}_n\omega^2 mg^2$$

This indicates a state of resonance when

$$\mathbf{T}_n + \mathbf{T} - \omega^2\mathbf{I} = 0 \quad . \quad . \quad . \quad . \quad (7)$$

made possible by the bending of the needle. This is seen to occur for a "loud" needle at 2 100 cycles per sec. in both Figs. 1 and 3. Substitution of the known values for $\mathbf{T}$, $\mathbf{I}$ and ω in equation (7) gives $\mathbf{T}_n$ the value 240×10^6 dyne-cm per radian.

At 30 cycles per sec. $\omega^2\mathbf{I} = 5{\cdot}4 \times 10^4$ and $\omega^2 mg^2 = 19{\cdot}2 \times 10^6$.

At this and lower frequencies we may neglect $\omega^2\mathbf{I}$ in comparison with $\mathbf{T}$, when y becomes

$$(\omega^2 mg^2 - \mathbf{T}_n)\mathbf{T} + \mathbf{T}_n\omega^2 mg^2 \quad \text{or} \quad (\mathbf{T} + \mathbf{T}_n)\omega^2 mg^2 - \mathbf{T}_n\mathbf{T}$$

When

$$\omega^2 mg^2 = \frac{\mathbf{T}_n\mathbf{T}}{\mathbf{T} + \mathbf{T}_n} \simeq \mathbf{T}_n \quad . \quad . \quad . \quad . \quad (8)$$

$y = 0$, indicating a state of resonance in the operation of the pick-up, the instrument vibrating as a whole about the point of the needle as axis. Such a state of vibration is known to exist at low frequencies, being made apparent by feeling the instrument with the tips of the fingers.

Substituting numerical values in the above expression, the resonant frequency is found to be 32 cycles per sec., which agrees with the value indicated in Figs. 1 and 3. This peak may readily be removed to a still lower frequency, if necessary, by increasing the mass of the pick-up or by attaching it rigidly to a massive supporting lever. The mass of the lever should, of course, be distributed in such a manner as to prevent excessive bearing pressure on the needle point.

At frequencies above about 3 000 cycles per sec. it is inadmissible to assume rigidity of the armature. The voltage output will fall off, moreover, owing to eddy-current effects in the magnetic circuit. The high-frequency peak may therefore serve the useful purpose of maintaining the voltage output at frequencies at which this would otherwise decrease. To achieve this it should occur at at least 3 000 cycles per sec.

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APPENDIX 3.

Tendency of the Pick-up Needle to Leave the Record Groove.

The profile of the groove in a record is approximately as shown in Fig. 6, and the forces exerted by the needle on the groove are illustrated there. If F becomes greater than mg the direction of the resultant force will subtend an angle of more than 45° with the vertical and the needle point will leave the groove. The tendency for this to happen is greatest at the lowest frequencies, when the amplitude of vibration is greatest. At such frequencies $\mathbf{I}d^2\theta/dt^2$ is small compared with $\mathbf{T}\theta$, so that the reaction force exerted by the needle point becomes

$$F = md^2x/dt^2 = \frac{\mathbf{T}\theta + Ad\theta/dt + \mathbf{T}_n\phi}{g}$$

or

$$F' = -\omega^2 mx' = \frac{\mathbf{T}\theta' + jwA\theta' + \mathbf{T}_n\phi'}{g}$$

At low frequencies the amplitudes of x and θ are high compared with ϕ, so that

$$F' \simeq \frac{\mathbf{T}\theta' + j\omega A\theta'}{g}$$

As previously shown, $y = x + g\theta$, hence

$$y' = -F'/(\omega^2 m) + g^2F'/(\mathbf{T} + j\omega A)$$

$$= F'\left(\frac{g^2\mathbf{T}}{\mathbf{T}^2 + \omega^2A^2} - \frac{1}{\omega^2 m} - \frac{g^2jA\omega}{\mathbf{T}^2 + \omega^2A^2}\right) = F'(a - jb) \quad (9)$$

It has been seen that if $F_{max.}$ exceeds m_1g * the needle will leave the groove, i.e.

$$y_{max.} \nearrow m_1g\sqrt{[a^2 + b^2]} \quad . \quad . \quad . \quad (10)$$

The tendency to leave the groove first becomes apparent, when using constant-power test-records, when the frequency is reduced to 100 cycles per sec. or lower. At this frequency we may neglect ω^2A^2 in comparison with $\mathbf{T}^2$ in equation (9), which reduces equation (10) to

$$y_{max.} \nearrow m_1g\left(\frac{g^2}{\mathbf{T}} - \frac{1}{\omega^2 m}\right) \quad . \quad . \quad . \quad (11)$$

Inserting numerical values, $y_{max.} = 0{\cdot}0205$ cm. The amplitude of the constant-power record of this frequency was 0·0037 cm, which is less than a fifth of the permissible value. At 50 cycles per sec. $y_{max.} = 0{\cdot}0131$ cm, while the groove amplitude was 0·0075 cm, or more than one-half of the permissible value. As might be expected from this, this particular pick-up followed the 50-cycle groove without jumping, although with considerable tone-arm vibration. Unfortunately, the values of g, $\mathbf{T}$ and m were not determined for the instruments which left the low-frequency grooves when under voltage test.

* It should be noted that the mass which provides the vertical force m_1g will not be the same as that which provides the horizontal force ω^2mx' in the case of a pick-up which is rigidly attached to a lever and counterbalance weight. m_1 will then be considerably less than m, owing to the effect of the counterbalance weight.

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FREQUENCY STABILIZATION OF VALVE OSCILLATORS.*

By Professor E. MALLETT, D.Sc., Member.

(*Paper first received* 8 *th December*, 1929, *and in final form* 4*th January*, 1930; *read before the* WIRELESS SECTION 5*th February*, 1930.)

SUMMARY.

The addition of a series inductance of suitable value to the anode circuit of a valve oscillator results in the frequency of the oscillations generated being independent of fluctuations in the valve constants.

There is no need to stress the importance of definitely fixing the frequency of transmitting stations, and the general practice to-day is towards using master oscillators of quite small power output to provide, through a series of amplifiers if necessary, the necessary grid swing of the power valves. Valve-maintained tuning-forks with frequency multipliers, and valve-maintained quartz crystals, are often used as the master oscillators on account of the great degree of frequency stability secured. The ordinary valve oscillator, however, has advantages, such as ease of frequency adjustment and simplicity of apparatus, not possessed by the mechanical oscillators, and, were it not for the frequency drift commonly encountered, it would undoubedly be far more used.

(1) THEORETICAL.

Considering the common type of oscillator in which an oscillatory circuit LRC is connected to the anode and a grid coil is coupled with L by a mutual inductance M, the frequency departures from the value $f = 1/\{2\pi\sqrt{(LC)}\}$ may conveniently be divided into two groups, primary and secondary. In the first group may be placed changes in the LC value itself, such as alteration of the effective value of L by a load coupled with L, and by grid currents flowing in the grid coil. The first of these is minimized by using the oscillator as a master, that is, by coupling to a second valve in which no grid current flows, and the second by using a bright-emitter valve showing definite saturation and having sufficiently high anode and grid biasing voltages, under which circumstances oscillations may be maintained without grid current flowing. These devices do not entirely avoid departures of this kind as there remain the grid-filament capacities through which currents flow, and the anode-filament capacity which shunts the condenser; but these are not likely to change.

The second group comprises the frequency departure imposed by the constants of the valve itself, and includes alterations due to change of anode- and filament-battery voltages, and of filament emission. The main effect of these changes on the frequency is through their alteration of the internal resistance R_a of the valve.

For an oscillator such as the one under consideration, neglecting the effects of valve capacities, the text-books find

$$\omega = \sqrt{\left[\frac{1}{CL}\left(1 + \frac{R}{R_a}\right)\right]} \quad . \quad . \quad . \quad (1)$$

for determining the frequency of the oscillations. Thus any change in R_a definitely affects the frequency. Suppose now that a series impedance Z' (inductance L', resistance R') is placed in the anode circuit as indicated in Fig. 1. Writing the parallel impedance of the LRC circuit Z_p and the series impedance Z, calling the anode current I_a and the current through L, I_1 (I_a and I_1 being vector currents), and supposing for a moment

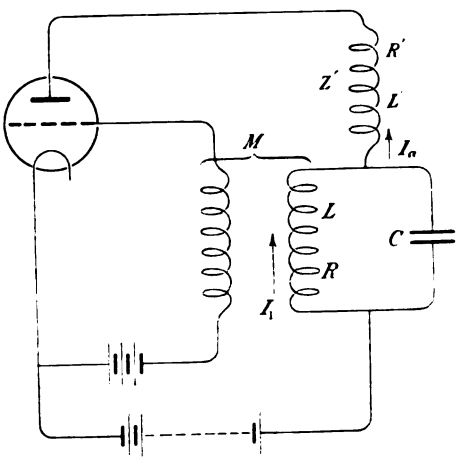

FIG. 1.

that a voltage E of frequency $\omega/(2\pi)$ is induced in the grid coil from an external source, then

$$I_a = \mu E/(R_a + Z' + Z_p) \quad . \quad . \quad . \quad (2)$$

and

$$I_1 = -\frac{j}{\omega C Z} I_a \quad . \quad . \quad . \quad . \quad (3)$$

and the grid voltage induced by I_1 is $V_g = j\omega M I_1 = M I_a/(CZ)$. Hence E in (2) must be replaced by $E + V_g$, giving

$$I_a = \frac{\mu}{R_a + Z' + Z_p}\left(E + \frac{M}{CZ} I_a\right)$$

i.e.

$$I_a\left(R_a + Z' + Z_p - \frac{\mu M}{CZ}\right) = \mu E$$

In order that oscillations may persist when E is made

* Reprinted from *Journal I.E.E.*, 1930, vol. 68, p. 578.

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zero, the left-hand side of this expression must also be zero, whence

$$R_a + Z' + Z_p - \frac{\mu M}{CZ} = 0 \quad . \quad . \quad . \quad (4)$$

Writing for Z, Z_p, and Z', their values, viz.

$$Z = R + j\left(\omega L - \frac{1}{\omega C}\right)$$

$$Z_p = (R + j\omega L)\left(-\frac{j}{\omega C}\right)\Big/\left\{R + j\left(\omega L - \frac{1}{\omega C}\right)\right\}$$

and $Z' = R' + j\omega L$

equation (4) gives

$$CR(R_a + R') - \omega CL'\left(\omega L - \frac{1}{\omega C}\right) + L - \mu M$$

$$+ j(R_a + R')\left(\omega L - \frac{1}{\omega C}\right)C + j\omega L'RC - j\frac{R}{\omega} = 0$$

Equating the imaginary terms to zero yields

$$\omega^2 = \frac{1}{LC} \cdot \frac{R_a + R + R'}{R_a + R' + (L'/L)R} \quad . \quad . \quad (5)$$

for the frequency.

Note that if $R' = L' = 0$, (5) becomes (1). Note also that if $L' = L$, (5) becomes

$$\omega^2 = 1/(LC) \quad . \quad . \quad . \quad . \quad . \quad . \quad (6)$$

and the frequency is quite independent of the valve constants.

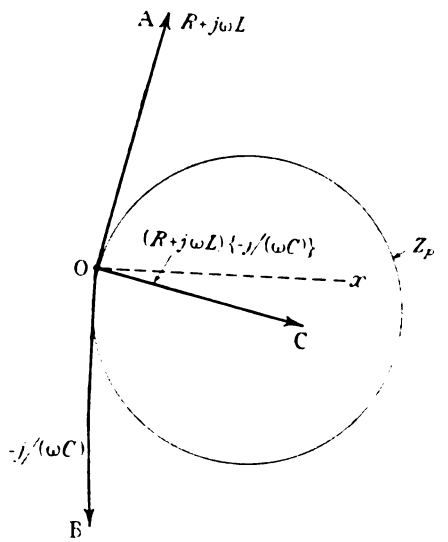

FIG. 2.

Equating the real terms to zero gives for the condition of maintenance

$$CR(R_a + R') - \omega CL'\left\{\omega L - 1/(\omega C)\right\} + L = \mu M$$

If $L' = R' = 0$ this becomes the well-known expression

$$CRR_a + L = \mu M$$

whereas if $L' = L$, so that $\omega = 1/(LC)$, it becomes

$$CR(R_a + R') + L = \mu M \quad . \quad . \quad . \quad (7)$$

These conclusions can be arrived at graphically in an illuminating manner. The condition for oscillations is, from (4),

$$R_a + Z + Z_p - \frac{\mu M}{CZ} = 0 \quad . \quad . \quad . \quad (8)$$

The locus of $1/Z$ is a circle with a horizontal diameter of length R, and the complex quantity $(R + j\omega L)\left\{-j/(\omega C)\right\}$ is found in Fig. 2, where OA $= (R + j\omega L)$, OB $= -j/(\omega C)$, and OC is the product of these two and varies very little with the frequency-changes involved.

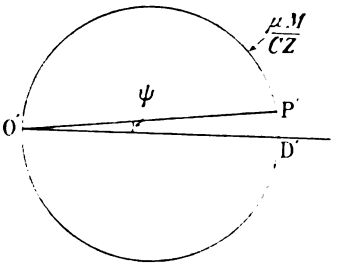

FIG. 3.

Thus the locus of Z_p is very nearly a circle drawn with its diameter in the direction OC; the diameter is the complex quantity $(R + j\omega L)\left\{-(\omega C)\right\}/R$.

Fig. 3 shows the locus of $\mu M/(CZ)$ — a circle with a horizontal diameter O'D' of length $\mu M/(CR)$.

The vector diagram corresponding to (8) is drawn in Fig. 4. OA is made equal to R_a, AA' to R', and $A'B$ to $j\omega L'$. From B the Z_p circle is drawn as in Fig. 2. To complete the diagram the subtraction of $\mu M/(CZ)$ must bring the point back to O, i.e. a point P must be found on the Z_p circle, such that OP $= \mu M/(CZ)$.

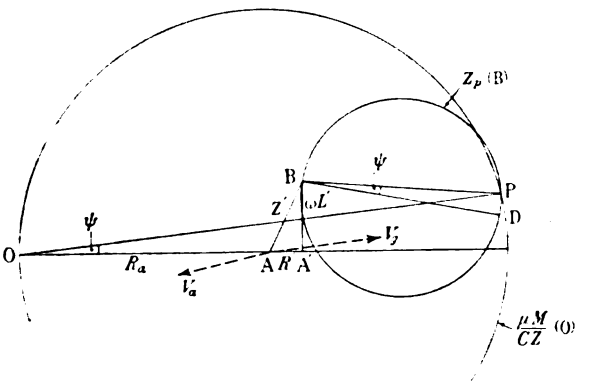

FIG. 4.

It is necessary therefore that O'P' in Fig. 3 drawn parallel to OP in Fig. 4 shall make the same angle ψ with the diameter O'D' that BP makes with BD in Fig. 4. The angle ψ determines the frequency, and the extent to which the circle of Fig. 3 must be enlarged to make O'P' = OP determines M. In the diagram the frequency is below resonance. If A'B (i.e. $\omega L'$) is so adjusted that D falls on OA produced, then the point P must fall on D, P' falls on D' and the frequency is that of resonance of the circuit, i.e. $\omega^2 = 1/(LC)$.

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Since $BD = (R + j\omega L)\left(-\frac{j}{\omega C}\right)/R = \frac{L}{CR} - j \cdot \frac{1}{\omega C}$ it is clear that D will fall on the axis when $A'B = 1/(\omega C)$; i.e. $\omega L' = 1/(\omega C)$ or $L' = 1/(\omega^2 C) = L$, and that then

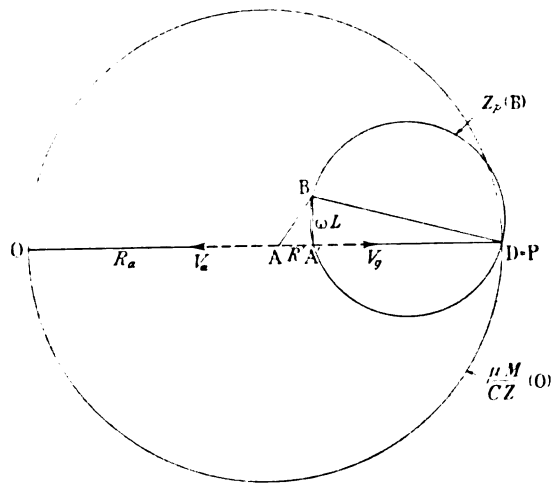

FIG. 5.

O'D' = OD, or $\mu M/(CR) = R_a + R' + L/(CR)$ in agreement with (7). The diagram is now as drawn in Fig. 5.

Since $V_a = -I_a(Z' + Z_p)$
$= -I_a(AB + BD) = -I_a \times AD$

V_a is drawn horizontally to the left from A.

And since $V_g = \frac{M}{CZ} I_a$
$= \frac{M}{CR} I_a$

V_g is drawn horizontally to the right from A. V_a and

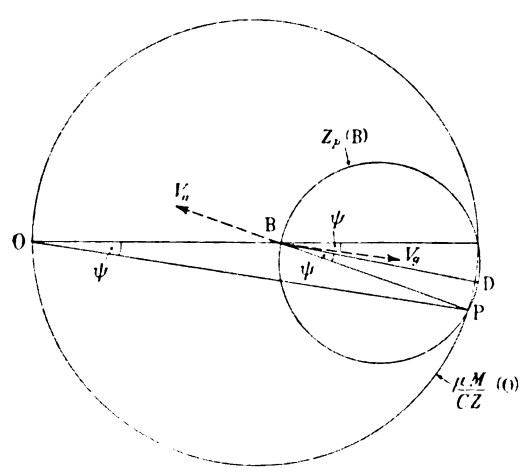

FIG. 6.

V_g are thus in exact phase opposition. With Z' omitted, that is in the usual circuit, the vector diagram is as drawn in Fig. 6, and the resonant frequency is higher than resonance. V_a is drawn in extension of PB and V_g is parallel to OP; there is not exact phase opposition. Similarly, in Fig. 4, V_a is drawn in the direction PA produced, and V_g is drawn parallel to OP. Here again there is not exact phase opposition.

The effect of adding the inductance L' of suitable value to the circuit is thus to bring the grid and anode voltages into exact phase opposition, which results—according to the simple theory given—in the frequency of the maintained oscillations being that of the free undamped oscillations of the oscillatory circuit.

(2) EXPERIMENTAL.

Some preliminary experiments, made with the assistance first of Mr. C. R. Urwin, and later of Mr. W. J. E. Tobin, have indicated that the possibility of frequency stabilization in the manner described does exist. The valve chosen was a Mullard ORA which, with an anode battery voltage of 300 volts and a grid-bias voltage of 34 volts, maintained oscillations in a circuit of inductance 350 μH and capacity 556 $\mu\mu$F without grid current flowing and without alteration of mean anode current. The characteristic curve with these

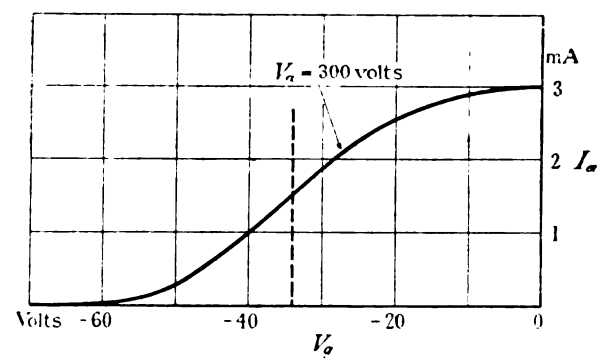

FIG. 7.

adjustments is given in Fig. 7. The frequency was thus about 360 000 cycles per sec. (wavelength 835 m).

A second oscillator was set up some distance away to "receive" the oscillations of the first by heterodyne, and a third (low-frequency) oscillator enabled the heterodyne note to be adjusted by beats to a fixed value by alteration of the frequency of the second oscillator. The small alterations of frequency required were made (and estimated) by alteration of the capacity of a standard 200-$\mu\mu$F condenser with vernier adjustment in series with a fixed 10-$\mu\mu$F condenser the combination being in parallel with the main oscillatory condenser. If C_1 is the capacity of the standard condenser and C_2 that of the fixed condenser (= 10 $\mu\mu$F) then the additional capacity across the oscillatory condenser is $C_1C_2/(C_1 + C_2)$, and if the change of this is δC, then $\delta f = \frac{f_0}{2C_0}\delta C$, where f_0 in the experiments was 360 000 cycles per sec., and C_0 was 556 $\mu\mu$F.

It appeared that the insertion of resistance in the oscillatory circuit lowered the frequency, instead of raising it as is indicated by equation (1). This was ascribed to alteration, by stray capacity effects, of the effective LC value, i.e. a primary alteration not due to

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the valve. This effect was great enough to mask any improvement due to the insertion of an inductance. It was, however, found that the insertion of the inductance L' did lower the frequency.

Changing the anode battery voltage yielded more promising results. It was found to be necessary to avoid cutting off the battery when changing over, and a quadruplex increment key was used (as indicated in Fig. 8), in which contact with the front stop is made before contact with the back stop is broken, the resistances " r " preventing excessive currents flowing while contact is made with both stops.

The results of one preliminary experiment are given

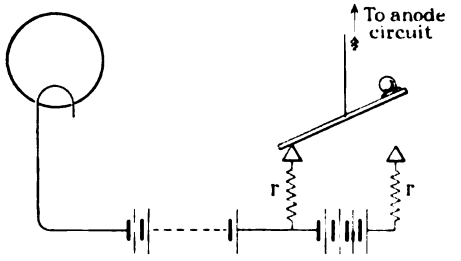

FIG. 8.

in the table. The added inductance L' is given in the first column. In the second and third columns are given the readings of the condenser C_1, A for an anode voltage of 270 volts and B for an anode voltage of 300 volts. In the fourth and fifth columns the corresponding values of $C_1C_2/(C_1 + C_2)$ are entered. These were read off from a chart prepared from the calibration curve of C_1. δC in the sixth column is the difference of these, and δf in the last column is the frequency change, reckoned positive if an increase of anode battery voltage causes an increase of frequency.

The frequency change produced by the 30-volt change in anode battery is plotted against the added inductance L' in Fig. 9, and from the curve obtained it would appear that with an added inductance of 63 μH no frequency-change would have been produced.

Altering the filament current by short-circuiting a small resistance in series yielded no positive result; the frequency alteration due to the change made was very small, and was not made any smaller by the insertion of inductances.

The fact that a smaller value of inserted inductance was found experimentally than is required by the theory is of interest. It may be due to the effect of the capacity between the anode and the grid, and the anode coil and the grid coil,* or to the effect of resistance in the capacity arm of the oscillatory circuit.† Other capacities may also be expected to modify the vector diagram; Appendix 1 is merely illustrative.

Other experiments at different frequencies have shown that the correct value for the inductance does depend upon the frequency and that the stray capacities may be such that the value of L' required is negative, i.e. a

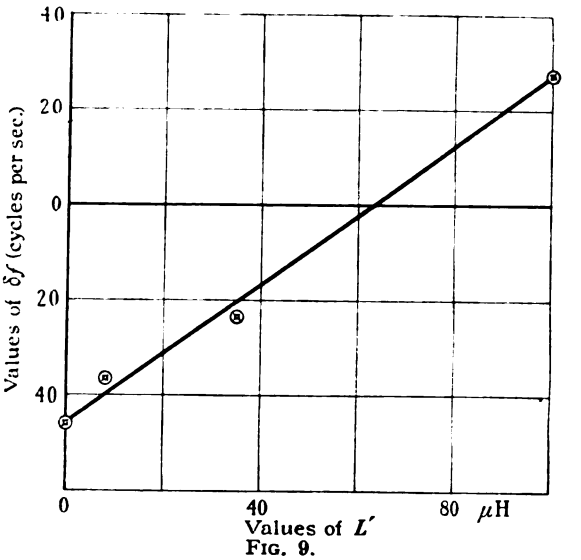

FIG. 9.

capacity must be added. Intermediately it may just happen that the circuit is already stabilized at the particular frequency in use.

Experiments are being continued with a view to ascertaining whether the triode circuit modified in this way can take the place of the tuning fork or quartz crystal. The mechanical vibrator is always placed in a thermostatic chamber to avoid frequency-changes due to temperature-changes; it would equally be necessary to place the whole of the oscillatory circuit of the electrical oscillator in a constant-temperature chamber.

While only one of many valve oscillator circuits (including those with mechanical vibrators) has been examined, it may be said in conclusion that the same principle would appear to apply to all—frequency stabilization is carried out by a circuit adjustment which eliminates the influence of the valve constants on the frequency —and that this adjustment will be found to ensure

Added inductance L'	Reading of C_1		$C_1C_2/(C_1 + C_2)$		δC	δf
	A	B	A	B		
μH			$\mu\mu F$	$\mu\mu F$	$\mu\mu F$	Cycles/sec.
0	94·8	119·0	9·20	9·34	0·14	— 45
8	69·2	78·3	8·98	9·09	0·11	— 36
35	89·2	98·0	9·16	9·23	0·07	— 23
100	61·5	56·7	8·90	8·82	— 0·08	26

* See Appendix 1. † See Appendix 2

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that the grid and anode potentials are in exact phase opposition.

The experimental work has been done in the Telegraph and Telephone Laboratories of the City and Guilds (Engineering) College, and the author wishes to thank Professor C. L. Fortescue, O.B.E., Member, for the interest that he has taken in the work.

APPENDIX 1.

Effect of Anode-Grid Capacity.

In Fig. 10 the condenser C_g represents the combined effect of the anode-grid capacity and the capacity between the anode and grid coils.

Let I' be the current through the grid coil and the

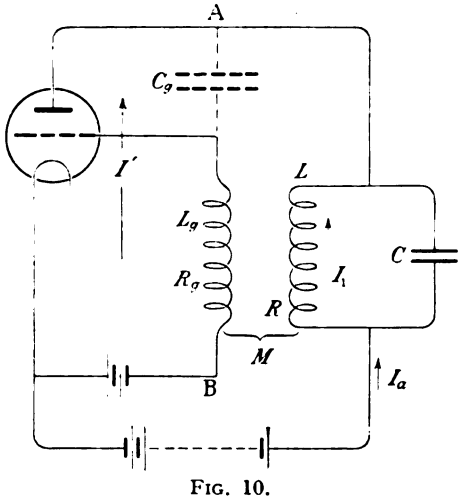

Fig. 10.

capacity C_g, and R_g be the resistance and L_g the inductance of the grid coil. The current I' is small compared with I_a, and is due mainly to the voltage V_a established across the points A and B, and only to a smaller degree to the voltage induced from the anode coil.

Thus, approximately,

$$I' = \frac{V_a}{R_g + j\omega L_g - j/(\omega C_g)} \doteqdot j\omega C_g V_a$$

$$V_a = -Z_p I_a$$

$$\therefore V_g = j\omega M I_1 + (R_g + j\omega L_g) I'$$

$$= \left\{ \frac{M}{CZ} - j\omega C_g Z_p (R_g + j\omega L_g) \right\} I_a$$

and the condition for oscillations becomes

$$R_a + Z_p + j\mu\omega C_g Z_p (R_g + j\omega L_g) - \frac{\mu M}{CZ} = 0$$

or

$$R_a + Z_p (1 - \mu\omega^2 L_g C_g + j\mu\omega C_g R_g) - \frac{\mu M}{CZ} = 0$$

The general effect on Fig. 6 is a reduction of the diameter BD and a decrease of the angle of depression. Thus a smaller inserted inductance L' is required to bring D on to the axis.

It might therefore be possible to effect the adjustment for stability by adding to the anode-grid capacity instead of by the insertion of the inductance L'.

APPENDIX 2.

Effect of Resistance in Capacity Arm of the Oscillatory Circuit.

Let the resistance of the capacity arm be R_c and of the inductance arm be R_l, and let $R = R_c + R_l$.

Equation (3) becomes

$$I_1 = \frac{R_c - j/(\omega C)}{Z} I_a$$

and equation (4)

$$R_a + Z' + Z_p - \frac{\mu M}{CZ} - \frac{j\omega\mu M R_c}{Z} = 0$$

with

$$Z_p = (R_l + j\omega L)\left\{ R_c - j/(\omega C) \right\}/Z$$

This yields for the frequency of the oscillations

$$\omega^2 = \frac{R_a + R' + R_l}{LC\left\{ R_a + R' + (L'/L)R + R_c - \mu M R_c/L \right\}}$$

and in order that this may be $1/(LC)$,

$$L' = L\left(\frac{R_l}{R} - \frac{R_c}{R} + \frac{\mu M R_c}{RL} \right)$$

For maintenance with greatest output power

$$CRR_a = L \text{ (nearly)} \quad \text{and} \quad \mu M = 2L.$$

The expression then becomes $L' = L$, as before. If, however, the inductance is too great (i.e. if an anode tap is required for largest output, which generally happens at high frequencies), the value of L' required is less than L, whereas if the inductance is too small, L' must be greater than L.

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AN INSTRUMENT FOR PROJECTING AND RECORDING THE RESPONSE CURVES OF ELECTRICAL CIRCUITS.*

By Professor C. L. FORTESCUE, O.B.E., M.A., Member, and F. RALPH.

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SUMMARY.

The paper describes a combination of a projection oscillograph, a special amplifier, and a variable condenser coupled to the rocking mirror of the oscillograph, by means of which response curves can be thrown on to a screen for observation or may be recorded photographically.

The amplifier may also be used with the oscillograph in the ordinary way for low-frequency observations, the amplifier then functioning as a special form of valve voltmeter giving an output suitable for operating the oscillograph.

(1) INTRODUCTION.

There are many occasions when it is desirable to observe with a minimum of trouble the response curves of electrical circuit arrangements, such, for example, as those of an ordinary wireless receiver, over some predetermined range of frequency. This can, of course, always be done by taking the necessary observations and plotting the curves, but where there are a large number of variables this becomes too long a process for the proper investigation of the properties of the circuits. The instrument described in this paper is an attempt to overcome this difficulty by a combination of a specially designed amplifier and a projection-type Duddell oscillograph. The horizontal deflection of the spot is made dependent upon the frequency change, and the vertical deflection is dependent upon the amplitude of the output of the circuits under test. The curve is repeated some 10 or 15 times per second, just sufficiently rapidly for the normal persistence of vision to give the impression of a continuous curve. This curve is projected on to a small screen in view of the operator adjusting the circuits. The effects of changing any particular variable are thus immediately seen. Photographic apparatus of the usual type can also be used for recording all response curves of sufficient interest to justify the time and expense involved.

In addition to this, the terminals are so arranged that the low-frequency parts of the instrument can be used for converting small voltages into currents large enough to operate the oscillograph. This is useful in such cases as the measurement of leakage fluxes where even an oscillograph current would cause appreciable changes in the flux being measured.

(2) THE SOURCE OF THE HIGH-FREQUENCY CURRENTS.

A variable-frequency screened oscillator is used in which the frequency is dependent upon the inductance and capacity in the oscillatory circuit. Both inductance and capacity are varied to give the necessary ranges of frequency change. The circuit arrangements are such that the wave-form is reasonably good over the whole range of frequency change, but at the same time the voltage amplitude in the oscillatory circuit remains constant. This adjustment is easily obtainable with any simple magnetically-coupled circuit, if the grid bias is so arranged that the change of anode current at the onset of oscillations is comparatively small. The coupling should be a little greater than the minimum value for which the oscillations will start.

The capacity in the oscillating circuit is in two parts. One part consists of a variable condenser adjusted by hand in order to give the correct mean value of the frequency—the appropriate inductance having been decided upon beforehand. The second part consists of a motor-driven variable condenser made up of two sets of fixed vanes and two sets of moving vanes, the latter being mounted on a long extension of the shaft of the driving motor. The two parts of the double condenser are in series and, as the vanes rotate, the capacity varies from maximum to minimum and back again during one revolution. This arrangement has the advantage that there are no sliding contacts in the high-frequency circuit. It has the disadvantage, however, that the moving vanes are not at earth potential, and consequently an insulating coupling must be used to connect the condenser spindle to the motor shaft.

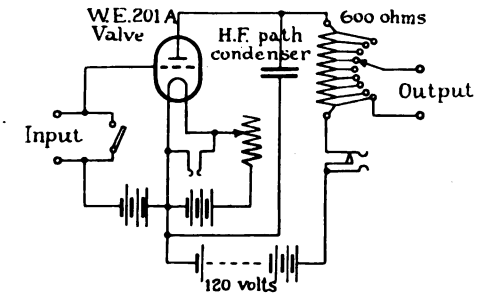

FIG. 1.

This motor-driven condenser is arranged to give a capacity of from about 50 to 200 $\mu\mu$F. By choosing various values of the inductance and hand-adjusted capacity, it is usually possible to arrange for this change of capacity to give the desired range of frequency. The only difficulty experienced occurred when a very large range of frequency was required.

In some cases it is more convenient to keep the frequency of the oscillator constant and to vary the tuning of the receiving circuit. This can be done, where the receiver consists of only a single resonant circuit, by

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connecting the motor-driven condenser in parallel with the capacity of this resonant circuit.

An extension of the condenser spindle carries a cam operating the usual rocking mirror of the oscillograph. By this means the spot of light is deflected horizontally in proportion to the change of capacity above and

resistance, and no more than 1 per cent of the rectified current passes through the capacity.

If a condenser were not used in this way the high-frequency components would go through all stages of the amplifier and upset the proportionality of the response to the rectified component. This necessity

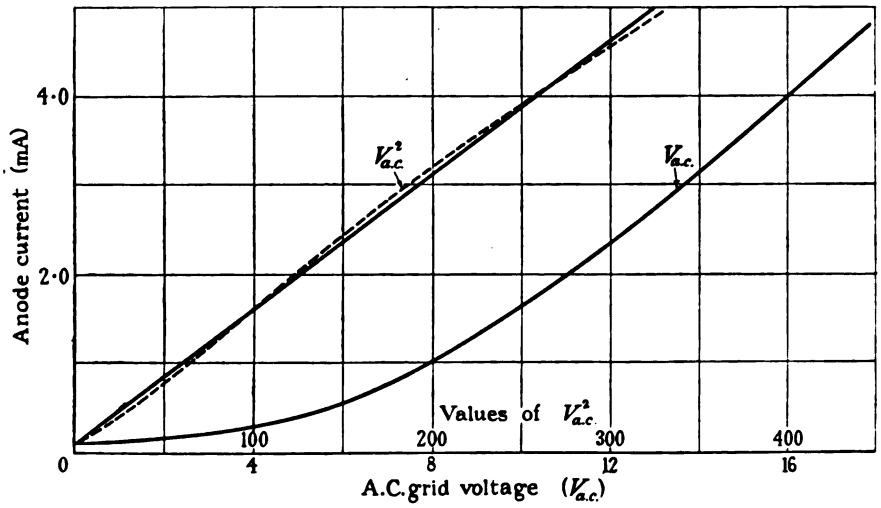

FIG. 2.

below the mean value. If this change is small compared with the total capacity, the horizontal scale is then practically a frequency scale. When desired, the

imposes certain limits on the frequencies, since the oscillation frequency must be sufficiently above the frequency of the cyclic variation for this separation to

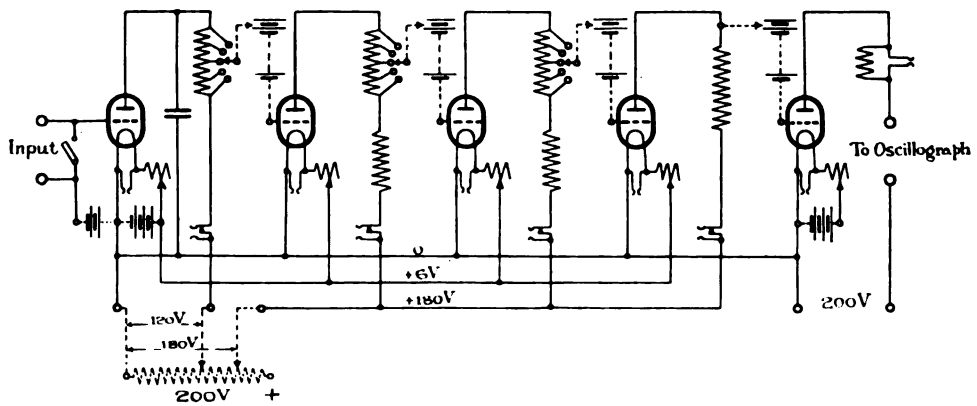

FIG. 3.

frequency scale can be checked at several points by means of a wavemeter.

The motor shaft also carries a shutter which cuts off the light for one half-revolution. This is necessary as the rotary condenser plates may not be exactly symmetrical.

(3) THE RECTIFYING UNIT.

This consists of a single Western Electric 201A valve connected as shown in Fig. 1. The output voltage from the circuit is applied directly to the grid, superimposed upon a steady grid-bias voltage of about 6 volts. The anode circuit consists of a resistance with a capacity across it so proportioned that no more than 1 per cent of the high-frequency component passes through the

be effected. The maximum values of the cyclic variation depend upon the harmonics of the wave-form of the response curve, and include all those that can be recorded by the oscillograph. Any harmonics higher than this need not be provided for as they will be lost in any case owing to the moving strips and mirror being unable to follow them. The maximum cyclic frequency may therefore be put at about 1 000. The minimum oscillation frequency is then in the neighbourhood of 100 000, depending upon the degree of accuracy which is required.

The Western Electric 201A valve was chosen as experience has shown that the rectification follows the square law over a considerable range. Fig. 2 gives the result of an actual test with one of these valves.

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(4) The Amplifier.

The output voltage of the rectifying unit may be as much as 1 volt. The R.M.S. current required by the projection-type oscillograph is about 0·15 ampere. Thus with a variation of input voltage of, say, $\pm$ 1·4 volts, the variation of the output current is to be about $\pm$ 0·2 ampere, and these two should be proportional to one another to within an accuracy of a few per cent.

When the amplifier is used without the rectifying unit for flux measurement, the voltages are often as low as 0·1 volt (R.M.S.). The amplifier is therefore designed to deal with only this small input if necessary, and to be used with larger inputs with a reduced amplification per stage.

A further special requirement in this case is that the amplifier should perform equally well at zero cyclic frequency, i.e. with step-by-step changes of input voltage. This is necessary for calibration purposes and immediately determines one point in the design of the instrument, viz. that it must be resistance-coupled with separate grid batteries for each stage.

The output stage was determined by the variation of the current required for the oscillograph. Eight Mullard D.F.A.7 valves working with an anode voltage of 200 and a mean anode current of about 175 mA were found satisfactory.

Three intermediate stages were employed as follows:—

1st stage: Mullard P.M.5B valve (V_a = 180 volts).

2nd and 3rd stages: One Mullard P.M.1 L.F. valve (V_a = 180 volts).

The resistances in the anode circuits were 84 000, 26 000 and 50 000 ohms respectively. Some variation of resistance was provided in the first two intermediate stages. Fig. 3 gives the connections for the complete amplifier. Other combinations of valves and resistances have also been used, with similar results.

With careful adjustment it was found possible to compensate for the curvature of the characteristics of the output valves by that of the valves in the intermediate stages. This is possible because each stage reverses the phase and a current limitation on one valve leads to a voltage limitation on the next in the direction in which the current would normally be too large.

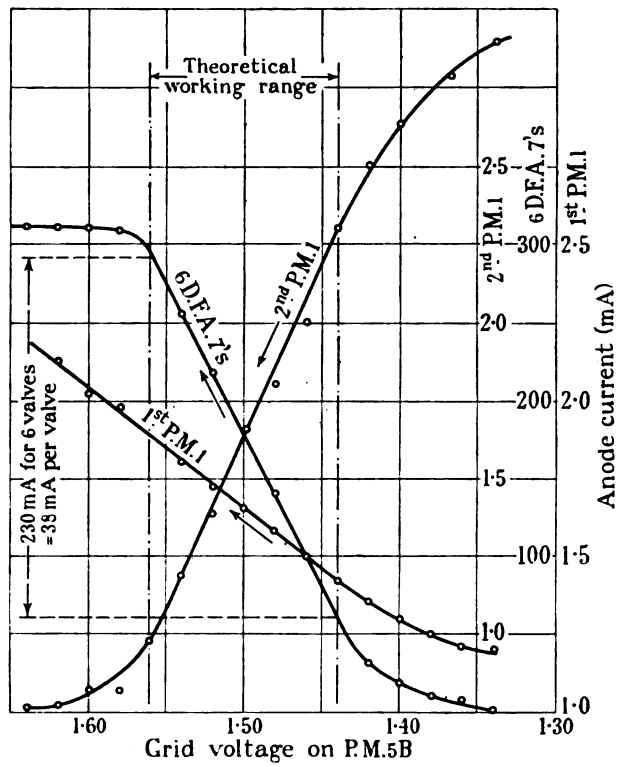

Fig. 4.

The final result for the whole amplifier, but with only six valves in the output stage, is shown in Fig. 4, where the current in the output circuit is shown plotted against the voltage applied to the grid of the first amplifier valve.

Very similar curves are obtained with eight valves in the output stage.

Where the maximum amplification is not required it is generally convenient to cut out the second amplifying valve by taking the grid connection of the third valve to the anode of the first valve instead of the second. The straight-line amplification is still preserved within the limits of accuracy required.

(5) The Oscillograph.

This was a standard Duddell instrument constructed by the Cambridge Instrument Co. The strips were

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larger than those fitted to the ordinary instruments, in order to provide for the larger mirrors necessary for projection purposes. The large strips were capable of carrying a considerable current, but actually the mean current is usually diverted by a battery and resistance in parallel with the strip. For this purpose the

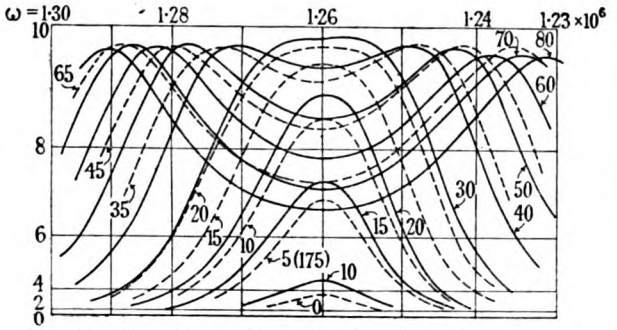

FIG. 5.—Test No. 3. Mutual only. Secondary current.

parallel resistance is made about 10 times the strip resistance so that about 90 per cent of the current change passes through the strips themselves.

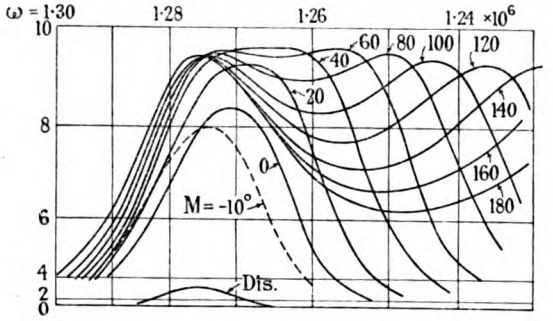

FIG. 6.—Test No. 9. Capacity coupling; $M = 0$. Secondary current.

(6) CIRCUIT ADJUSTMENTS.

The mean anode currents for all the stages were found by preliminary experiments and kept constant throughout. The amplifier was fitted with a voltmeter and a milliammeter, which could be connected in turn to each stage to verify the conditions. Once adjusted, the only changes called for were those due to batteries running down. This is perhaps the most serious difficulty in using the instrument, and it can only be overcome by using good batteries throughout and keeping them in good condition.

(7) EXAMPLES OF THE APPLICATION OF THE AMPLIFIER.

(*a*) *Response curves.*—The curves of Figs. 5 and 6 are reproduced directly from records obtained from

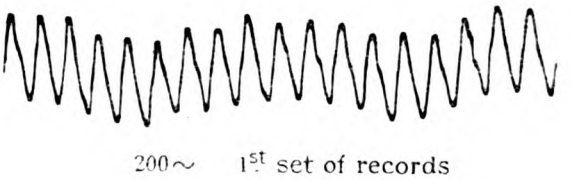

FIG. 7.

coupled circuits, magnetic and capacity coupling respectively.

(*b*) *Other applications.*—Fig. 7 is reproduced from a record of the output of an electrical gramophone pick-up when working from a " sine wave " record. The output is not sinusoidal and the question arose as to whether this was attributable to the amplifier, the pick-up or

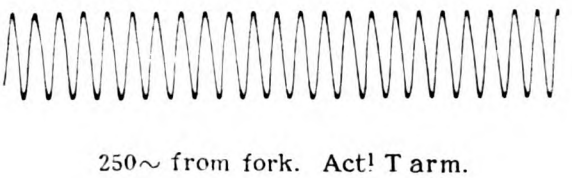

FIG. 8.

the record. On driving the pick-up by means of a maintained fork the curves of Fig. 8 were obtained; they are obviously much more nearly sinusoidal, thus showing that the irregularities of Fig. 7 are due to the record itself.

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THE DEVELOPMENT OF A PRECISION AMMETER FOR VERY HIGH FREQUENCIES.*

By E. B. MOULLIN, M.A., Associate Member.

(*Paper first received 6th November, 1929, and in final form 15th January, 1930.*)

SUMMARY.

This paper describes the principles and development of an ammeter which has been made for measuring current of very high frequencies, such as 30 000 kilocycles per sec. The instrument is of the dynamometer type and has a geometrical form for which all the changes of current distribution can be calculated. The correcting factors approach a limiting value and do not increase continuously with frequency as they do in thermal instruments. A special feature of the design is that an instrument can be made to carry unshunted a current of any magnitude. This removes all uncertainty about shunting ratio and so the instrument should be more reliable than one in which the current is supposed to divide evenly among a number of resistance wires in parallel.

Methods are described for checking the calculated correction factors and there are also some calibrations at frequencies round about 1 000 kilocycles per sec. In addition, measurements are described of the high-frequency resistance and of the inductance and terminal capacity.

(1) INTRODUCTION.

The difficulty of measuring currents of very high frequency has again been made prominent by the development of short-wave telegraphy. The instruments in use at present are no doubt relatively correct at any one frequency, but possibly they may have an appreciable frequency error. Thermal instruments, with or without a current transformer, have been used almost exclusively for measuring radio-frequency currents, and, since the resistance of all conductors increases continuously with frequency, all such instruments must become progressively more sensitive as the frequency rises. The heated element is made from a wire the diameter and material of which are such as to make the probable increase of resistance small at the highest frequency required, but when frequencies of the order of 10^8 cycles per sec. are contemplated the wire must be so fine that it can scarcely carry a current of 1 ampere. Many wires must be placed in parallel, and then it is difficult to ensure that the fraction carried by each one is independent of frequency. To this end, usually the wires are arranged as generators of a surface of revolution, for example a cylinder, hyperboloid or double cone. Given perfection of manufacture this device must achieve the required result in respect of the effect of one wire on the others, but it affords no protection from a magnetic field penetrating the whole cage. If the cage is placed concentrically within a long and thick screen tube of copper, then the current will divide evenly between all the wires at all frequencies and in all circumstances. The instrument will still have a frequency error due to the increase of resistance of each wire, but this should be calculable with a high degree of accuracy. It will have a further frequency error due to capacity effects, and this may not be readily calculable.

The author has been concerned in making an ammeter which can carry a current of any magnitude unshunted and which has a geometrical form for which the various frequency corrections can be calculated exactly. Uncertainty in respect of unconstant shunting ratio is then removed entirely and the frequency corrections are known and are, in fact, very small. This ammeter has been described already † and the analysis of the various correction factors developed in great detail. Since then considerable experience has been gained in the construction and use of the instrument, and this paper is concerned chiefly with such recent developments. It will be necessary first to describe briefly the principle of the instrument and to enumerate the cause and magnitude of the three separate frequency corrections, after which the constructional details and methods of testing will be given.

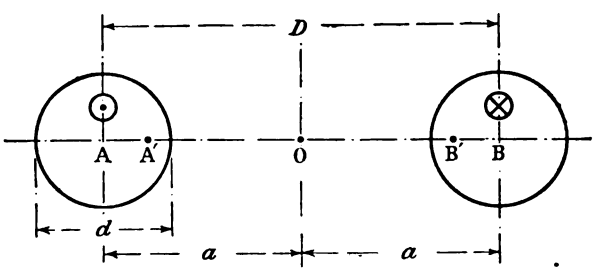

FIG. 1.

(2) ELEMENTARY SYSTEM OF THE AMMETER.

Consider two long, straight, solid circular cylinders of diameter d, placed parallel to one another at a distance D apart, and let them carry the same current in opposite directions; the system is shown diagrammatically in section in Fig. 1. The two cylinders will repel one another with a force $F = 2i^2/D$ per unit length if the current i is distributed uniformly over the cross-section. If the current is alternating and of R.M.S. value I, the mean value of the repulsive force is $F = 2I^2/D$. If the force can be measured in any convenient manner the system becomes an ammeter which can be calibrated

* Reprinted from *Journal I.E.E.*, 1930, vol. 68, p. 544

† *Proceedings of the Royal Society*, A, 1928, vol. 121, p. 41.

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with steady current and used to measure alternating current. But an alternating current is not distributed uniformly over the cross-section, and the force is no longer given accurately by the equation $F = 2I^2/D$ and the instrument requires a frequency correction. In the limiting condition of infinite frequency or perfect conductivity there is no magnetic field within the cylinders and then each circumference must coincide with a line of force. It is well known that each circumference would coincide with a line of force of two line currents situated at the points A′ and B′ which are mutually inverse points for the circles centred at A and B. So ultimately we have

$$F = \frac{2I^2}{A'B'} = \frac{2I^2}{\sqrt{(D^2 - d^2)}} \quad . \quad . \quad . \quad (1)$$

We see that the ultimate increase of force is small so long as D is considerably greater than d.

A frequency correction which comes to a limiting value is in striking contrast with the frequency correction for thermal ammeters, which does not come to a limiting value but increases continuously. Furthermore, the limiting correction is sensibly attained at finite frequencies. Very complicated analysis is required to express the force as a function of frequency, but the result is expressed by the following series:—

TABLE 1.

D/d	F_0	F	F/F_0
$1\frac{1}{2}$	1·296	1·340	1·032
2	1·13	1·15	1·02
3	1·048	1·06	1·01
4	1·026	1·031	1·005

$$F = \frac{2I^2}{D}\left\{1 + \frac{1}{2}\frac{d^2}{D^2} + \frac{3}{8}\frac{d^4}{D^4} - \frac{1}{\sqrt{2z}}\cdot\frac{d^2}{D^2}\left(1 + \frac{d^2}{D^2}\right) + \ldots\right\} \text{ with } z > 3{\cdot}5\ldots \quad (2)$$

where we have the well-known parameter $z^2 = \pi\omega d^2/\rho$, in which $\omega = 2\pi f$ and ρ is the resistivity of the material. Substitution in (2) will show that the limiting value has sensibly been attained when z is, say, greater than 10, and this is illustrated by Table 1, in which F_0 is the force for zero frequency.

So if $D/d = 3$ the limiting increase is less than 6 per cent and this has been attained within about 1 per cent of itself when $z = 10$. This will occur at frequencies less than 300 kilocycles per sec. for all copper wires larger than No. 16 S.W.G.

But the two cylinders, as depicted in Fig. 1, will inevitably be situated in a magnetic field arising from the remainder of the circuit, and this field will produce an additional force. Therefore the wires must be shielded from external fields by placing them within a thick copper tube in the manner shown diagrammatically by Fig. 2. This tube will prevent the penetration of any external field, but the currents situated at A and B will induce in the screen tube eddy currents whose field will modify the force on the wires. The screen tube is comparable with the short-circuited secondary of a transformer, and so the currents in it are substantially in antiphase with the currents at A and B. The field between A and B due to the eddy currents in the screen is therefore opposed to the field created by the currents at A and B and consequently lessens the force on each. If the small non-uniformity of current density round the circumferences of the cylinders at A and B is now ignored, it follows readily that in the limit the field of the eddy currents within the screen tube is the same as would be the field of image currents placed at A′ and

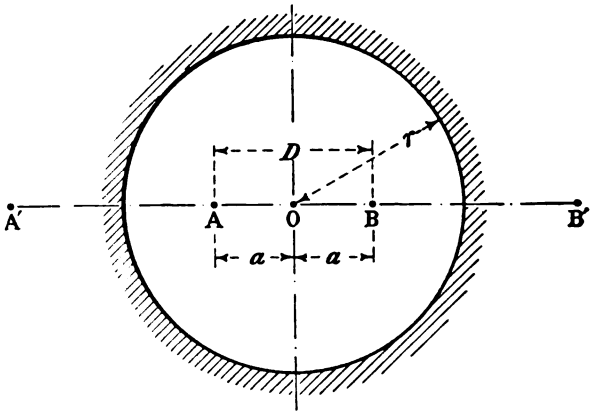

FIG. 2.

B′, the inverse points of A and B with respect to the circle of radius r centred at O. It follows from this that

$$F = \frac{2I^2}{D}\left(1 - \frac{4OA \,.\, OA'}{A'B \,.\, B'B}\right) = \frac{2I^2}{D}\left(1 - \frac{4r^2a^2}{r^4 - a^4}\right) \quad . \quad . \quad . \quad . \quad (3)$$

Since the eddy currents in the screen reduce the force between the cylinders, they tend to counteract the increase of force due to non-uniform current distribution round the cylinders. We may superpose the two effects without sensible error (see Appendix), and since $r >> a$ and $D > d$ the two effects will neutralize each other approximately when a, r, etc., are so chosen that $D^2 = rd/\sqrt{2}$. If $r/a = 8$ (approx.), then in the limit either effect by itself would change the force due to a specified current by about 6 per cent, or the current required for a given force would differ by 3 per cent from that required at zero frequency, since the force varies as the current squared. The difference between two approximately equal effects, each comparatively small in value, must be a quantity of negligible magnitude. The effect of the screen approaches close to its limiting value at a frequency of a few hundred cycles per second, but this may be considered more appropriately later.

We now have to consider the effect of longitudinal distribution, which is important only when the length

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of the system is comparable with the wavelength. Two parallel cylinders which are not in a screen tube may be represented by the circuit system shown diagrammatically by Fig. 3, which represents a uniform short-circuited line of length l and having a capacity and inductance per unit length of C and L respectively. If the current entering at A is $-I\cos\omega t$, and that

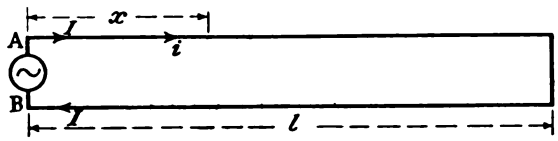

FIG. 3.

leaving at B is $I\cos\omega t$, it is well known that the current i at a distance x along the line is given by the equation

$$i = -I\frac{\cos\omega a(l-x)}{\cos\omega al}\cos\omega t$$

where $a = \sqrt{(LC)}$. The voltage between the ends of the line is expressed by the equation

$$v = \frac{I}{2}\sqrt{\frac{L}{C}}\frac{\sin\omega a(l-x)}{\cos\omega al}\sin\omega t$$

(The resistance of the line is ignored, but this makes only a second-order difference to the current distribution.)

If the surrounding medium has unit permeability and dielectric constant ϵ, and if c is the ratio of the electromagnetic to the electrostatic absolute unit of current or charge, then for a parallel-wire system we have $LC = \epsilon/c^2$. The force per unit length between the cylinders is a repulsion of value $2i^2/\sqrt{(D^2-d^2)}$ due to the magnetic field and an attraction of value $2q^2/\epsilon\sqrt{(D^2-d^2)}$ due to the electrostatic field. Hence for the electrostatic attraction F_e we have

$$F_e = 2\cdot\frac{4C^2v^2C^2}{\epsilon\sqrt{(D^2-d^2)}} = \frac{2I^2}{\sqrt{(D^2-d^2)}}\frac{\sin^2\omega a(l-x)}{\cos\omega al}\sin^2\omega t$$

and so for the mean total force we have

$$F = \frac{I^2}{\sqrt{(D^2-d^2)}}\int_0^l \frac{\cos^2\omega a(l-x)-\sin^2\omega a(l-x)}{\cos^2\omega al}dx$$

$$= \frac{I^2}{\sqrt{(D^2-d^2)}}l\frac{\tan\omega al}{\omega al} \quad . \quad . \quad . \quad . \quad . \quad (4)$$

$$= \frac{I^2l}{\sqrt{(D^2-d^2)}}(1+\tfrac{1}{3}\omega^2a^2l^2+\tfrac{2}{15}\omega^4a^4l^4)$$

approximately, if $\omega al \ll 1$. (4a)

Equation (4a) may be interpreted more readily in terms of wavelength than of frequency. When $\pi al = \frac{1}{2}\pi$ the length of the system equals a quarter wavelength of current; since the system is immersed in a dielectric of constant ϵ the length of the current waves on it are thereby made $1/\sqrt{\epsilon}$ of the length they would have in a dielectric of constant unity. Let the wavelength in free space, corresponding to a frequency $f = \omega/(2\pi)$, be λ. Then we may rewrite (4a) as follows:—

$$F = \frac{I^2l}{\sqrt{(D^2-d^2)}}\left(1+\frac{4}{3}\cdot\frac{\pi^2\epsilon l^2}{\lambda^2}+\frac{32}{15}\cdot\frac{\pi^4\epsilon^2l^4}{\lambda^4}\right) \quad . \quad (5)$$

So if $\epsilon = 2$ and $l/\lambda = 1/25$ the force is 4·5 per cent greater for a given current than it is at zero frequency. We are contemplating a system in which l is about 20 cm, and then a given current will produce 2·2 per cent more deflection at a wavelength of 5 m than it did at zero frequency. The correction term is therefore quite definite and calculable but it is so small as to be unimportant for the wavelengths in use at present.

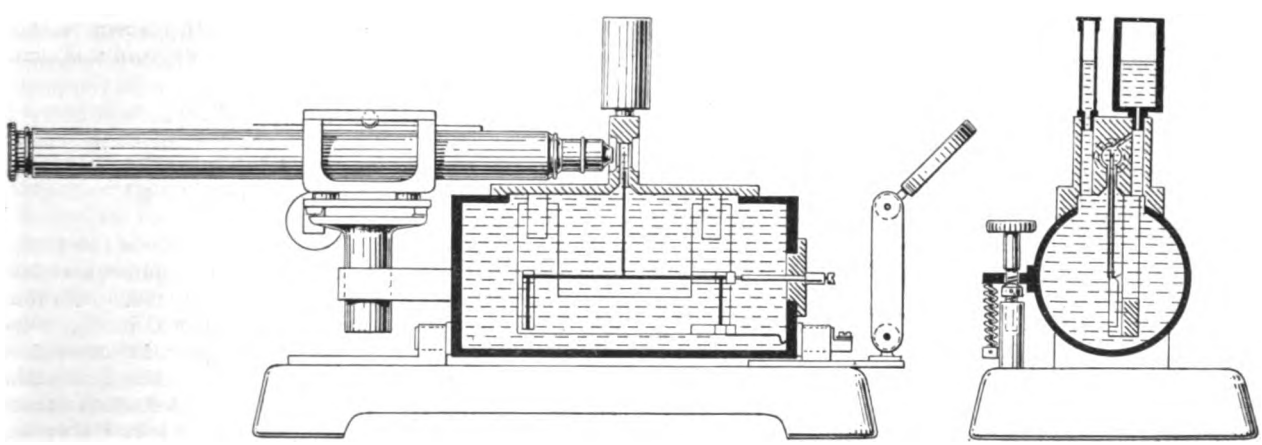

FIG. 4.

This analysis is incorrect for a system enclosed in a screen tube. The correct analysis has been developed,* but it is very complicated and it is unnecessary to reproduce it in this paper. When the system is in a screen tube the point B in Fig. 3 will be connected to the screen, and then the complete analysis shows that the correcting factor within the screen differs insensibly from that calculated from (5), provided l is replaced by $2l$. So if the instrument has been arranged with $D^2 = rd/\sqrt{2}$,

* *Proceedings of the Royal Society*, loc. cit.

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the only resultant correcting factor is that due to longitudinal distribution of current: for a system 20 cm long this will not have reached 2·5 per cent even when the wavelength is reduced to 10 m.

(3) Constructional Details of the Ammeter.

So far only two of these ammeters have been made, the first for currents up to about 2 amperes and the second for currents up to about 25 amperes. A longitudinal and transverse section of the first instrument is shown in Fig. 4 and two longitudinal sections of the second instrument in Figs. 5A and 5B. These drawings show clearly the arrangement of the cylindrical copper screen tube, which in both instruments is 3 mm thick and 77 mm internal diameter and 190 mm long. One cylinder is fixed rigidly to an ebonite frame attached to the under side of the thick copper cover-plate of the screen tube. The other cylinder stands on spring legs which conduct the current to it and also provide most of the elastic constraint to resist the repulsive force between the two cylinders. The repulsive force of the current moves the two cylinders further apart and it is this spread which provides the indication of the ammeter. The repulsive force is very small, about $2I^2/10^4$ grammes weight in the instrument of Fig. 5, and this necessitates the spring legs being weak and the total spread very small. The spread is measured by observing with a microscope the movement of a thin quartz pointer attached to the middle of the moving cylinder. This pointer passes through a hole in the cover plate of the screen and is viewed through a glass window in a small chamber. The arrangement of pointer, window and microscope may be seen in Figs. 4 and 5 and requires no further explanation. The pointer tip magnifies the movement of the cylinder about five times; the microscope has a 1-inch objective with a 10-inch tube length and an eyepiece scale 1 cm long divided into 100 parts. Consequently the maximum spread of the cylinders is about 0·2 mm, and this is 2 per cent of their central distance for the instrument of Fig. 5. If the spread is x, then the law relating x to the current I is

$$I = A\sqrt{x}(1 - \tfrac{1}{2}x/D),$$

and this differs inappreciably from a simple square law since the greatest value of x/D is only 2/100. Also it is clear from equations (1) and (3) that if D increases by 2 per cent there is a negligible change in the correction factor. In the instrument of Fig. 5 the correction factor would change from zero at no deflection factor to + 0·25 per cent at full scale. The spring legs are made from standard cadmium-copper alloy, and this is a great improvement on the phosphor-bronze used originally in the small instrument. The cadmium-copper strip is 3 mm wide and 0·08 mm thick in the 25-ampere instrument and 0·5 × 0·035 mm in the 2-ampere instrument. The current density in the strips is very high, about 11×10^3 amperes per cm^2, and for this reason the whole screen tube is filled with a high-grade oil. Besides helping to cool the strips, the oil damps the movement and makes the instrument dead-beat and at the same time makes a cushion for the movement when the instrument is carried about. The 2-ampere instrument would be very fragile without

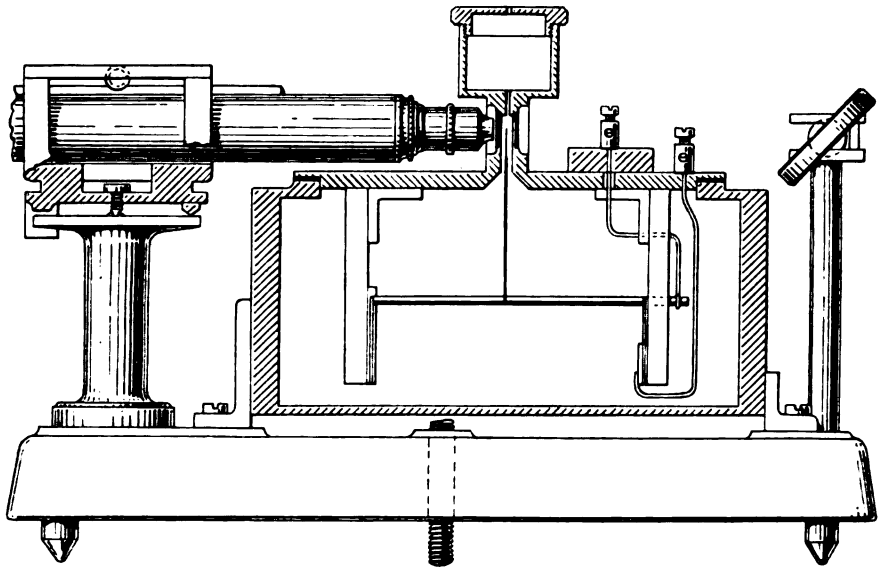

Fig. 5A.

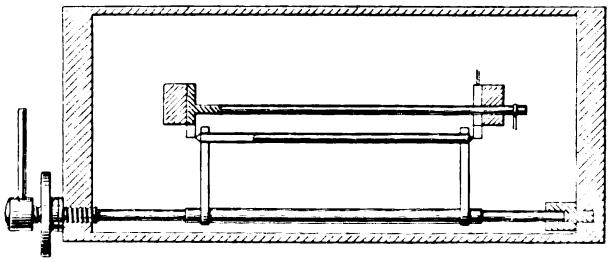

Fig. 5B.

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the oil; but with the oil it is perfectly portable and has withstood several long journeys standing unpacked on the floor of a light motor-car. The 25-ampere instrument is thoroughly robust and would not need oil to make it portable; the comparatively heavy cylinder renders a clamp desirable and its arrangement is shown in Fig. 5B. Two light phosphor-bronze fingers are fixed to the clamp rod, and a quarter-turn of this lifts the fingers so that the cylinder is cradled in forks at their ends. It will be noticed that the end bearing of the clamp rod is of ebonite, and this prevents a flow of induced current along the rod and back by the screen.

The terminal arrangement differs from that illustrated in Fig. 4 and will be discussed later.

The oil immersion is useful also for reducing the load carried by the spring legs, and in the 2-ampere instrument these are in tension. At the same time it reduces sensitivity as it reduces the gravity control, which control is negative when the net weight is positive.

In the small-current instrument the moving cylinder is an aluminium tube 10 cm long and having an external diameter of 1·1 mm with walls 0·025 mm thick. In the large-current instrument it is a tube of copper-alloy (verging on brass) 10 cm long, 3·19 mm external diameter and having walls 0·4 mm thick and weighing about 2·5 grammes in oil.

The force of gravity is used to place the cylinders at the exact central distance obtaining at the time of testing and calibration. A fixed fiducial pointer is mounted inside the pointer chamber, and the microscope, which has a short cross-traverse, is placed so as to bring the fixed pointer exactly at one end of the eyepiece scale. The levelling screw, shown in Figs. 4 and 5, is then turned until the gravity control brings the moving pointer to the other end of the scale; this setting is taken as standard. Having set the pointers, the microscope is further cross-traversed until the fixed pointer has moved out of the field of view and the moving pointer stands at the zero of the eyepiece scale. The width of the pointer chamber is such that the central distance of the cylinders cannot change by more than 0·6 mm in 10 mm, and experiment with the larger instrument shows that a given current can produce an extreme range of deflection of ± 5 per cent above and below the mean and ± 1 per cent in the smaller instrument. The setting-up process therefore does not require to be done with very great care.

It was found that a deflection due to heating was present in the 25-ampere instrument (but not in the 2-ampere instrument). This was traced to imperfect parallelism of the spring legs. The mounting has been fitted with an adjustment which twists slightly the bottom of one spring. The slow drift deflection due to heat effect can be made positive or negative by this adjustment, which is so set as to make it zero before the instrument is sealed up. Experience shows that this adjustment remains permanent. In addition to this there may possibly be a slow drift of about 2 mm due to convection currents in the oil. Apparently this can often be neutralized by slight adjustment of the levelling screw. However, searching tests show that if the deflection is observed and the zero is read at the instant of switching off, a given current produces the same deflection consistently to within less than ± 0·3 mm in 100 mm. This means that full-scale deflection represents a given current to within ± 0·15 per cent, and so the instrument justifies the correction factors which are to be applied to it.

The pointer chamber is fitted to take a condensing lens so that the image of the pointer may readily be thrown on a screen. The instrument is perhaps more convenient to use in this manner and there is a large gain in the obtainable accuracy of reading.

The pointer chamber is surmounted by a cylindrical reservoir to form a breathing chamber for the oil.

(4) Calibrating and Testing the Ammeter.

It is not easy to set up the system on its spring legs so that the cylinders are at precisely the desired distance apart, and it is desirable to check this distance after the ammeter has been sealed up in its case. This may be done by placing parallel currents at external points such as A′ and B′ in Fig. 2. Thus, referring to Fig. 2, let there be a winding of N turns with its sides at A′ and B′ and connected in series with the ammeter in the sense to reduce the deflection. By a suitable choice of N and the distances AA′, BB′ it is easy to make the deflection almost zero for any current. A permanent magnet should be placed near the ammeter to neutralize the vertical component of the earth's magnetic field. The method of adjustment is to approach the magnet until there is no change of deflection on reversing the direction of a steady current of given value. (Note that during this adjustment the external winding is cut out.) If this adjustment is imperfect, so that a small remaining field V produces a difference of deflection which is a fraction y of the mean deflection due to a current I, then we have $VD/(2I) = y$. Now let the external winding be put in series with the ammeter, having one coil side at a distance b from A, in Fig. 2, and the other at a distance b'. Also let the deflection due to I be now a fraction x of its value without the external winding, then it follows readily that

$$D = \frac{(1+x)(1+y)bb'}{N(b+b')} = \frac{bb'(1+x+y)}{N(b+b')}$$

since both x and y are very small. So by measuring b and b', D may be deduced. Naturally the system will be arranged with b nearly equal to b', and then it is worthy of notice that a given fractional error in measuring b makes a very much smaller fractional error in the value deduced for D. So D may be deduced to about the same accuracy that $(b + b')$ can be measured. As an example of such a test in the 25-ampere instrument, $x = 2·9$ per cent, $y = 0·8$ per cent, $bb' = 18·4$ cm, $b = 6·8$ cm and $N = 4$, whence we find $D = 1·11$ cm. The internal diameter of the screen tube was 77 mm, and so it follows from (3) that the limiting reduction of deflection due to the presence of the screen should be $4(11·1/77)^2 \times 1/100$ per cent = 8 per cent. The cylinders are a little too far apart for perfect compensation, since the relation $D^2 = \frac{1}{2}Rd$ would require that $D = 0·935$ cm. So the two limiting correcting factors

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do not balance by 4·15 per cent and in the limit the ammeter would read 2·07 per cent lower than the steady-current calibration.

q is every odd integer. Taking $t = 0{\cdot}3$ cm, $r = 4$ cm and $\rho = 1{\cdot}68$ microhms per cm^3, we find $m = 0{\cdot}028\,f$. On substitution it will be found that $m^2 = 197$ when the

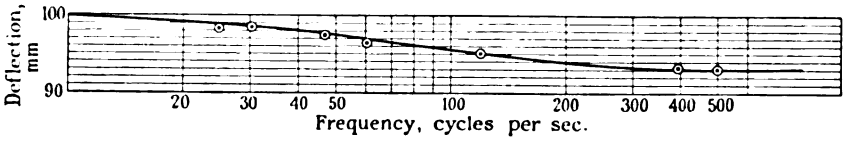

FIG. 6.

Let us now consider the rate at which the effect of the screen approaches the limiting value. If the non-uniformity of current density across the thickness t of

frequency is 500 cycles per sec., so that at this frequency the screen correction will have reached its limiting value to within 0·5 per cent. Over this range it is a

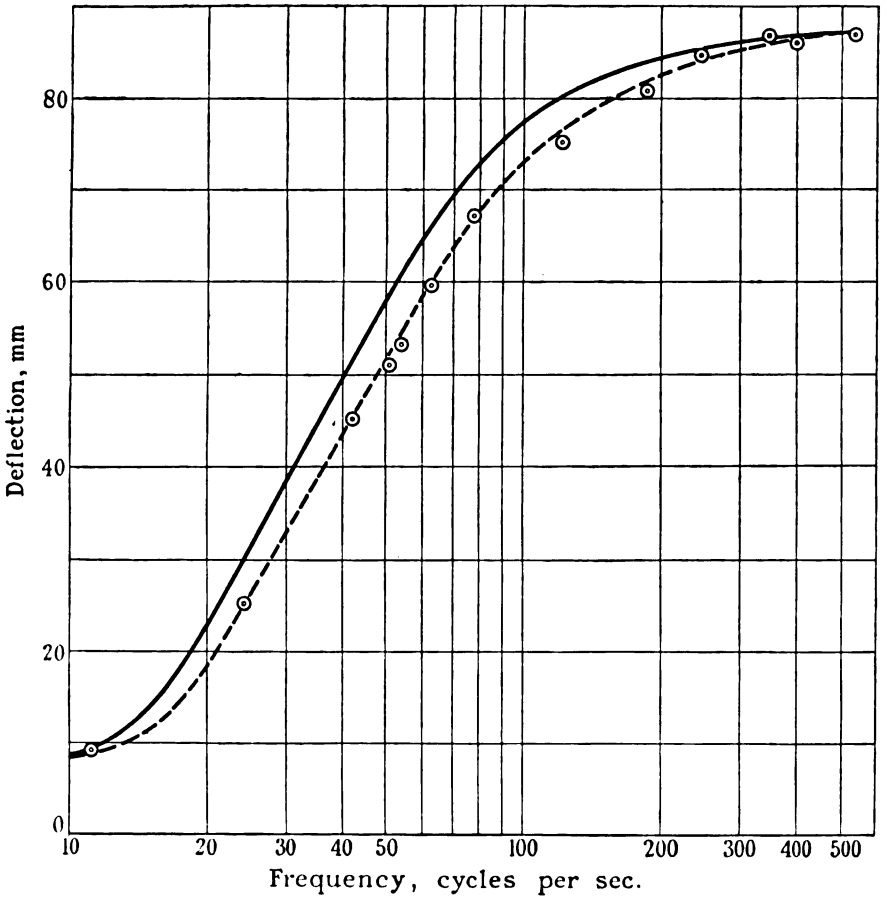

FIG. 7.

the screen tube is ignored, it may be shown * that equation (3) should be written

$$F = \frac{2I^2}{D}\left\{1 - 4\sum_1^\infty \left(\frac{a^2}{r^2}\right)\frac{qm^2}{m^2+q^2}\right\} \quad . \quad . \quad (6)$$

$$\doteqdot \frac{2I^2}{D}\left\{1 - 4\frac{a^2}{r^2}\cdot\frac{m^2}{m^2+1}\right\} \text{ since } \frac{a^2}{r^2} << 1$$

where $m = 2\pi t\omega r/\rho$, in which $\omega = 2\pi \times$ frequency, ρ is the resistivity of the tube whose mean radius is r, and

simple matter to plot experimentally a curve connecting the deflection due to a given current and the frequency; in this way the accuracy of the predicted correcting factor can be tested. This has been done for the 25-ampere instrument and the result is illustrated in Fig. 6, and from this the limiting reduction appears to be 6·6 per cent, which is 1·7 per cent less than was anticipated. These tests took a considerable time to make and are obviously very difficult since the differences are so small. It was thought that a more reliable method of making the test would be to switch over instantly from steady

* *Proceedings of the Royal Society*, loc. cit., p. 53; also F. W. CARTER: *Proceedings of the Cambridge Philosophical Society*, 1927, vol. 23, p. 902.

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current to 500 cycles per sec. Three readings were taken with the steady current in each direction and with current at 500 cycles per sec., and the greatest departure from the mean was 0·5 mm. The deflections are as follows: steady current 97·5 mm, steady current reverse 96·8 mm, mean 97·15 mm; at 500 cycles 89·5 mm. According to this the reduction of sensitivity is 7·9 per cent, and this agrees with the predicted value of 8·3 per cent to well within the limits of experimental error.

The calculated correcting factor might be inexact because the analysis assumes that the cylinders and the screen tube are infinitely long, whereas the length of the cylinders is in reality only $2\frac{1}{2}$ times the radius of the screen. A large portion of the screen tube, at its ends, is not acted on by magnetic field and this might well be expected to reduce the eddy currents flowing. An indirect method of checking the correction factor is to see whether the deflection with the external coil in use varies in the correct manner with frequency. With steady current the external field neutralizes the deflection, but as the frequency increases less and less of this field penetrates the screen and the deflection will tend to the value it has with no external coil. This experiment will surely be an exacting test of the applicability of analysis appropriate to very long cylinders. If, in the previous notation, we take $b = b'$ it may be shown that

$$F\Big/\frac{2I^2}{D} = \frac{m^2}{m^2+1}\left[\frac{x}{m^2} + \left(1 - \frac{4a^2}{r^2}\right)\right]$$

A certain current was found to give a deflection of 87·2 mm at 560 cycles per sec. and 2·5 mm at zero frequency. The curve of Fig. 7 has been drawn from values calculated from the above equation, in which $\left[1 - (4a^2/r^2)\right]$ has been given the value 87·2 and x the value 0·03; the observed deflections at various frequencies are shown by observation points near the dotted curve. Experiment shows that the external field which penetrates the screen is a little more than the calculated value; this is doubtless due to the resistance of the copper end-pieces. However, the observed and calculated deflections are in substantial agreement, and the fractional difference between the two curves must represent very closely the fractional error in the calculated correction factor at a given frequency. Fig. 7 is surely good evidence to prove that the calculated limiting values of the correction factors are substantially accurate, and that if the two factors are arranged to neutralize approximately then the error in the calculated value of their small difference must be quite unimportant. If a curve corresponding to Fig. 7 could be obtained for the cylinders, presumably it would fit the calculated values still more closely, for there the end effect is relatively much less important. That one of the correcting factors can readily be measured experimentally, as shown by Fig. 6, is perhaps a feature of the instrument.

The central distance between the cylinders of the smaller instrument was deduced in the same manner. With a compensating winding of 6 turns the net deflection was negative and with 5 turns it was positive, and from the observed values of these deflections the central distance was calculated to be 8·7 mm and 8·65 mm respectively. For this distance the screen should reduce the deflection by 4·7 per cent. The steady-current deflection was deduced from the mean of three readings between three-quarters and full scale, and likewise the deflection for currents at 500 cycles per sec. The reduction of deflection at 500 cycles per sec. was thus found to be 4·9 per cent, and this again verifies the method of calculating the correction factor. The correction factor for cylinders 1·1 mm diam. and 8·7 mm apart has a limiting value of 0·7 per cent, and will have approached closely to this value at a frequency of 1 megacycle per sec.*

(5) Resistance, Inductance and Capacitance of the Ammeter.

(*a*) *Estimated values.*

The steady-current resistance of the larger instrument is 14·0 milliohms, which is composed as follows:—

	mΩ
Lead from earth terminal to moving cylinder ..	1·0
Lead from fixed cylinder to terminal	0·6
Moving cylinder	1·6
Fixed cylinder	0·2
The two spring-legs	10
	13·4

The fixed cylinder is solid copper and the moving cylinder is a tube of resistivity 6·2 microhms per cm³. The total length of circuit from terminal to terminal is 43 cm. The power input at full scale is about 7·5 watts at low frequencies. About 70 per cent of the total resistance is in the spring legs, which are too thin for their resistance to increase appreciably at any frequency likely to be used. Thus, supposing the current to penetrate into them as though they were plates of infinite width, then the current density at mid-thickness is still 10 per cent of its surface value at a frequency of 3×10^7 cycles per sec. ($\lambda = 10$ m) and 47 per cent at $\lambda = 50$ m. The increase of total resistance should, therefore, be mainly due to the cylinders and leads.

It may be shown that the resistance per unit length of each cylinder enclosed in a screen tube, as depicted in Fig. 2, may be calculated from the equation

$$R_n/R_0 = \frac{\sqrt{2z}}{4}\left\{\frac{D}{\sqrt{(D^2-d^2)}} + \frac{D^2rd}{2(r^4-a^4)}\right\}$$
$$\doteqdot \frac{\sqrt{2z}}{4}\left(1 + \frac{1}{2}\frac{d^2}{D^2} + \frac{1}{2}\frac{d}{r}\cdot\frac{D^2}{r^2}\right)$$

This equation shows that the presence of the screen and the proximity of one cylinder to the other cause a negligible increase of resistance; accordingly these effects will be ignored. Assuming that the resistance of the strips does not increase appreciably, then the resistance of the ammeter should be about 42 milliohms at $\lambda = 50$ m and 64 milliohms at $\lambda = 15$ m. At this wavelength the case will have to radiate about 0·4 watt per square inch of cooling surface.

* *Proceedings of the Royal Society*, loc. cit., Table 3.

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The steady-current resistance of the smaller instrument is 83 mΩ and is composed as follows:—

	mΩ
Moving cylinder	33
Fixed cylinder	1·5
The two spring-legs	47

The fixed cylinder is solid No. 18 S.W.G. copper, and the moving cylinder is an aluminium tube 1·1 mm diameter and 0·025 mm thick. The total length of the circuit is about 43 cm and the power input at full scale is about 0·4 watt at low frequencies.

If the effect of the screen is ignored, the limiting value of the inductance of the cylinders is

$$L = 4 \log_e \frac{D + \sqrt{(D^2 - d^2)}}{d} \text{ per unit length.}$$

If the effect of non-uniform current distribution round the circumferences of the cylinders is ignored, then the inductance in the screen is

$$L \doteqdot 4 \log_e \frac{2D}{d}\left(1 - \frac{a^2}{r^2}\right)$$

Combining these two we have as a first approximation

$$L = 4 \log_e \frac{2D}{d}\left(1 - \frac{1}{4}\frac{d^2}{D^2} - \frac{a^2}{r^2}\right)$$

and the correction term, as part of a logarithm, is evidently negligible. Taking $D = 1$ and $d = 0{\cdot}3$ we have $L = 7{\cdot}28 \times 10^{-3}$ μH per cm, or, taking $D = 0{\cdot}5$, then $L = 5 \times 10^{-3}$ μH per cm, so L is insensitive to the ratio D/d. At low frequencies, when the current is distributed uniformly, we should have respectively $8{\cdot}28 \times 10^{-3}$ and $6{\cdot}28 \times 10^{-3}$ μH per cm. Thus for any ammeter of this type the inductance will be roughly 7×10^{-3} μH per cm of cylinder. The change from cylinder to spring leg or terminal lead will evidently make very little difference and we shall estimate the inductance at 7×10^{-3} μH per cm of the system. Accordingly either instrument should have an inductance somewhere between 0·15 and 0·17 μH. Similarly the capacity of the system, if on open circuit at the far end of the loop, would be about 3·2 μμF, and twice this amount when immersed in oil.

Reduction from the exact equation * shows that the apparent input impedance is

$$Z \doteqdot \omega L l\left\{1 + \frac{8\pi^2 l^2}{3\lambda^2} \cdot \frac{2\alpha - 1}{\alpha - 1}\right\}$$

where l is the length of the loop, λ is $1/\sqrt{\epsilon}$ times the wavelength in the ether and α is a constant such that $\alpha = (\log_e 2r/d)/(\log_e r/D)$; numerical substitution will show that α has a value not far removed from 2. Alternatively this equation may be interpreted as showing the apparent input impedance to be due to the inductance shunted by a capacitance of value $\frac{1}{3}Cl(2\alpha - 1)/(2\alpha + 1)$, where C is the capacity per unit length in oil. Accordingly we must consider the ammeter to be shunted by a capacity of about 2·5 μμF, and in parallel with this capacity is the capacity due to the terminals external to the case.

* *Proceedings of the Royal Society*, loc. cit., equation (15).

(*b*) *Measured values.*

The resistance of both ammeters has been measured at various wavelengths between 50 and 150 m. The process was to connect a coil, whose apparent inductance was either 1 μH or 2 μH, in parallel with a variable condenser and to deduce the effective resistance of this system by plotting a resonance curve. The ammeter was then added to the circuit and the effective resistance again determined, the difference being ascribed to the ammeter. The earth terminal of the ammeter was in both measurements connected to the earth terminal of the condenser, and the ammeter was included by moving one end of the inductance from this terminal to the other terminal of the ammeter. The inherent resistance of the circuit was about 0·2 ohm over the range of wavelength used. Though the operation of including the ammeter makes a negligible disturbance to the coil, yet it is not necessarily correct to ascribe the difference of resistance to the ammeter because the capacity has to be reduced to compensate for the inductance of the ammeter. When the ammeter is not included, then R_1, the total apparent resistance, is composed of R_c, the effective resistance of the condenser, and R_L, the effective resistance of the coil. After including the ammeter we have

$$R_2 = R_c' + R_L' + R_A$$

It is fair to assume $R_L' = R_L$, but it is probable that R_c' is greater than R_c because the capacity has been decreased. So R_A is probably less than $R_2 - R_1$. Some values determined with the 2-μH coil are shown collected in Table 1.

TABLE 1.

λ	165 m	115 m	80 m	50 m
	ohm	ohm	ohm	ohm
R_1	0·153	0·18	0·285	0·344
R_2	0·213	0·266	0·373	0·444
$R_2 - R_1$	0·06	0·086	0·088	0·10

Now though these differences increase roughly as $\sqrt{f}$, yet they are much larger than the estimated resistances of the ammeter. For example, at $\lambda = 50$ m the apparent resistance is 100 milliohms instead of about 42 milliohms.

Measurements with the 1-μH coil are given in Table 2.

TABLE 2.

λ	80 m	50 m
	ohm	ohm
R_1	0·194	0·258
R_2	0·31	0·342
$R_2 - R_1$	0·116	0·084

The values by difference are in rough agreement with those found with the 2-μH coil, but consistency of repetition in the measurements showed that the agree-

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ment ought to be closer if the difference does represent the resistance of the ammeter.

From a different set of experiments, to be described at some future time, there was very strong reason to suppose that the condenser had an effective resistance of 0·16 ohm and 0·12 ohm at wavelengths of 50 m and 80 m respectively, at the capacity in use, with the 1-μH coil, before adding the ammeter. After adding the ammeter the capacity had to be reduced to 82 per cent of its former value; if the condenser had a constant power factor its effective resistance would then be 0·19 ohm and 0·145 ohm respectively. On this assumption the ammeter resistance would be respectively 0·06 ohm and 0·09 ohm at 50-m and 80-m wavelengths. It is clear, therefore, that the condenser can make a large effect on the value deduced for the resistance. But probably we can say that the high-frequency resistance is certainly not more than twice the estimated value. The

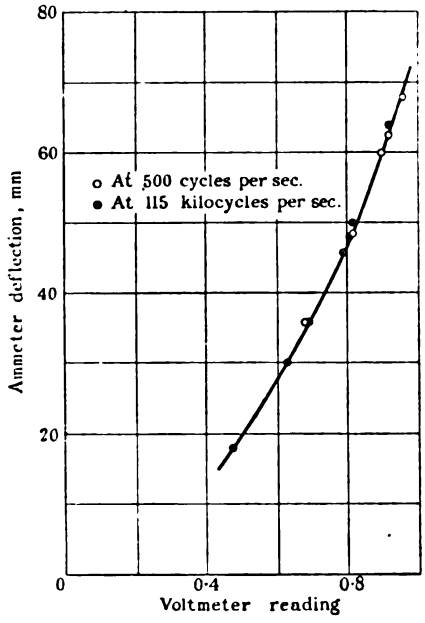

FIG. 8.

resistance of the small ammeter was measured at 50-m and at 80-m wavelengths, using the 2-μH coil, and found to be 0·187 ohm and 0·193 ohm respectively, showing an apparent increase of about 2·3 times the steady-current value.

The inductance of both instruments has been measured by difference, using the same two coils. With a given coil the inductance values by difference are consistent over a wide range of wavelength, but depend slightly on the coil used. The large ammeter has an inductance of 0·18 μH when measured by difference with the 1-μH coil, and 0·22 μH when measured with the 2-μH coil; these values are about 20 per cent greater than the estimated value. The inductance of the small instrument appeared to be 0·2 μH.

The capacity of the system has not been measured, because to do so would entail opening the system at the far end. The terminal capacity of both forms of terminal system has been measured. When measuring the capacity of the axial terminal, as in Fig. 4, the movement had been removed, but 4 cm of lead protruded into the case from the terminal. The capacity was measured by noting the necessary decrease of a variable condenser required to compensate for joining a neighbouring wire to the axial terminal, the capacity formed between the terminal and its lead and the case then being in parallel with the variable condenser. With the case full of oil, this capacity was found to be 3·6 $\mu\mu$F over a range of wavelengths, and 2·9 $\mu\mu$F with the case empty of oil. The capacity of the lead, considered as a piece of concentric cylinder, is 0·6 $\mu\mu$F in air and 1·2 $\mu\mu$F in oil. Subtracting these values we find the true terminal capacity to be 2·4 or 2·3 $\mu\mu$F.

The capacity of the terminal system shown in Fig. 5 was measured when the terminal lead was just freed from the fixed cylinder. The capacity in oil was found to be 4·1 $\mu\mu$F and this includes the capacity of 6 cm of terminal lead whose capacity belongs to the system. Again estimating this capacity as for a concentric

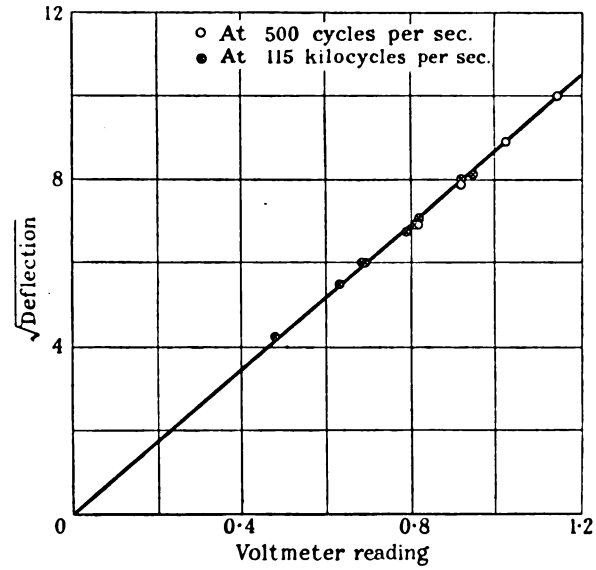

FIG. 9.

cylinder, we must subtract 1·8 $\mu\mu$F and again find a residue of 2·3 $\mu\mu$F due to the terminal proper. It therefore seems safe to assume that the true terminal capacity, shunting the system, is less than 3 $\mu\mu$F.

(6) TESTS AT HIGH FREQUENCY AND TEMPERATURE-RISE.

The smaller instrument has been checked very carefully at various frequencies up to 3×10^6 cycles per sec. Three 1-ampere hot-wire instruments (with 8-in. dials) were joined in parallel and then connected in series with the ammeter being tested. The relation between deflection and the sum of the three ammeter readings at 500 cycles per sec. was obtained with great care, and this has been compared with the relation obtained at radio frequencies. Also the voltage-drop across 0·39 ohm of No. 26 S.W.G. Eureka wire was measured by means of a thermionic voltmeter. The silk-covered Eureka wire was twisted on itself, making a very narrow

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loop 7 cm long, and the calculated inductance of this loop to steady currents is 0·03 μH. At 115 kilocycles per sec. the resistance will have increased by less than 1 part in 1 million and the reactance will increase the voltage-drop by 0·17 per cent. Tests were made with the voltmeter at 115 (λ = 2 600 m), and with the hot-wire ammeters at 1 850 (λ = 160 m), 2 300 (λ = 130 m) and 2 830 (λ = 106 m) kilocycles per sec. Fig. 8 shows the comparison between voltmeter reading and deflection: the points marked with a circle were obtained at 500 cycles per sec. and the points marked with a cross in a circle were obtained at 115 kilocycles per sec. Fig. 9 shows the voltmeter reading plotted against the square root of the deflection, and the average agreement between the points obtained at 500 cycles per sec. and those at 115 kilocycles per sec. is certainly closer than 1 per cent. Fig. 10 shows the sum of the three ammeter readings plotted against the deflection: the points marked with a dot in a circle were obtained at 500 cycles per sec. These high-frequency observations could not be made in very ideal circumstances, both because conditions were not perfectly steady at the very high frequencies, and also because sufficient current for full-scale deflection could not be obtained. At the frequencies to which Fig. 10 relates the proximity of the cylinders should increase the deflection for a given current by 0·7 per cent. Fig. 10 is satisfactory in showing that a given deflection always corresponds to a larger reading of the hot-wire ammeters at very high frequencies than at low frequencies, and so shows that the hot-wire instruments are consistently reading high: their error is, however, small and is of the order of 3 per cent round about λ = 100 m.

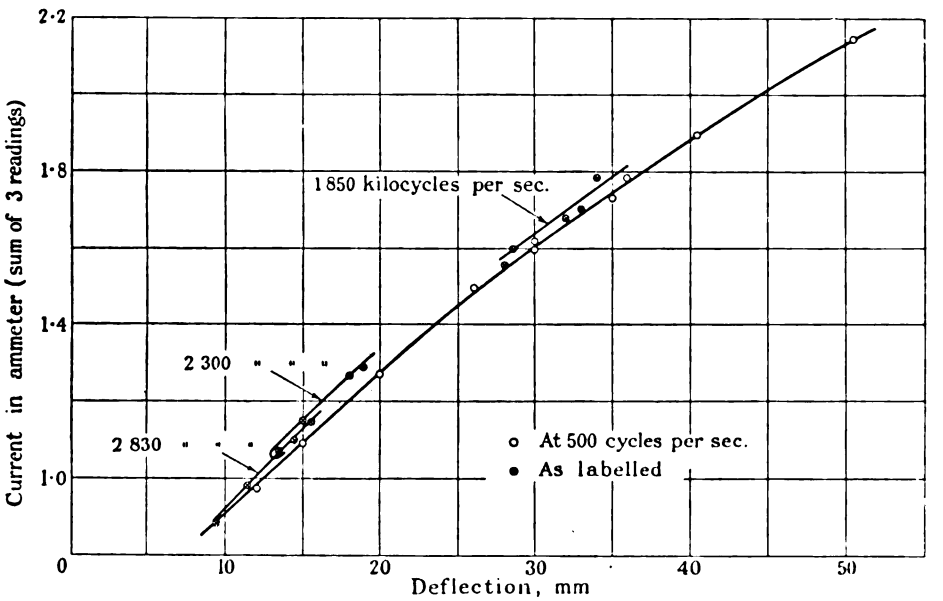

FIG. 10.

The large ammeter has been calibrated at 500 cycles per sec. against a Weston model 425 radio-frequency thermo-ammeter (0 to 20 amperes) and the two instruments again compared at 3 000 kilocycles per sec. (λ = 100 m). The result of the comparison is exhibited in Fig. 11, the points marked with a cross being obtained at 3 000 kilocycles per sec. Here again it was difficult to obtain perfectly steady conditions at the high frequency, and the current could be adjusted only in fixed steps which did not correspond to the most desirable parts of the scale. Since both instruments had a square-law scale, the four readings between 17 and 18 amperes should be the most reliable, and according to these a current at 3 000 kilocycles per sec. produces 5·6 per cent more deflection than it did at 500 cycles per sec. But now we have to allow for the proximity effect of two cylinders, 3·19 mm diam., placed 11·1 mm apart, and from formula (1) this will increase the deflection by 4·1 per cent. We have accordingly to account for a residue of only 1·5 per cent, and thus the scale reading of the Weston instrument is only 0·75 per cent high at a wavelength of 100 m—surely a very satisfactory performance for so small a thermal instrument of 20-ampere capacity.

Curves of temperature-rise for the large instrument are shown in Fig. 12, and by means of a cooling curve the thermal time-constant of the ammeter was found to be about 3 000 seconds. The heat capacity is about 4 600 joules and the cooling coefficient 1·5 watts per deg. C.

From the two heating curves of Fig. 12 we deduce that the resistance at λ = 100 m is 72 milliohms, and this value is in fair agreement with the values collected in Table 1.

(7) Calculated Correction Tables for the Large Ammeter.

These factors are expressed as a correction to be applied after the proximity effect has attained its limiting value. Thus we suppose the calibration has been made at some convenient frequency such as 200 cycles per sec.

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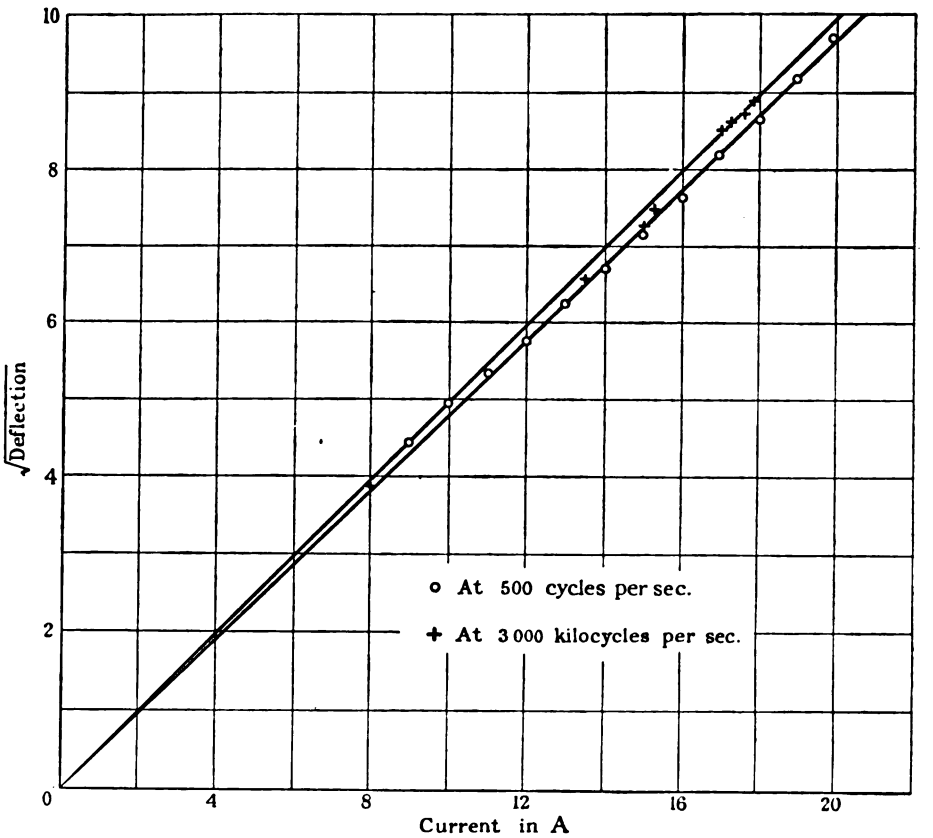

FIG. 11.

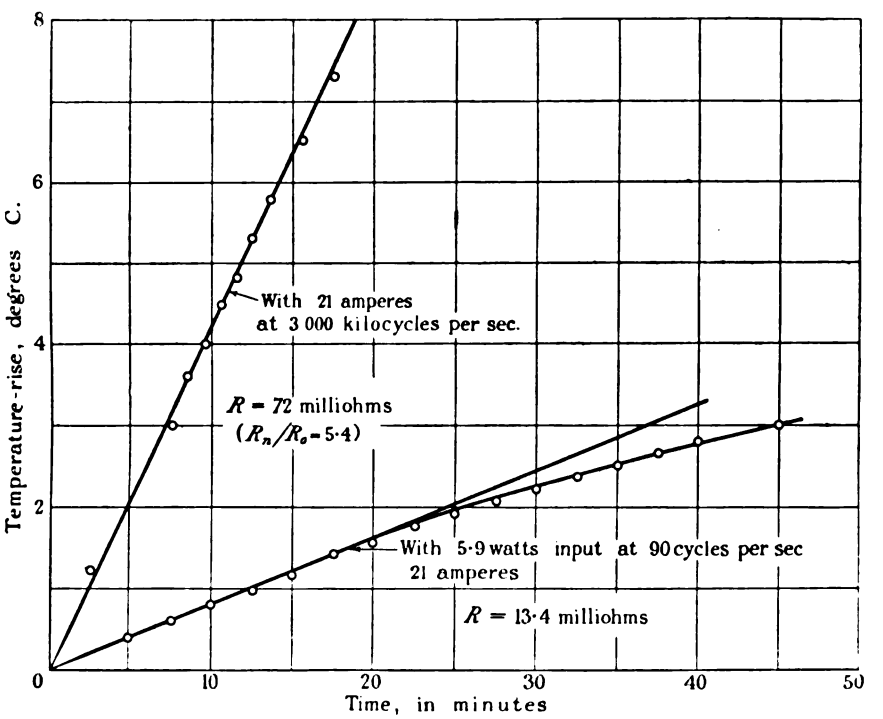

FIG. 12.—Heating curves of large ammeter.

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and that the deflections then observed have been increased by 4·1 per cent to render the calibration valid for frequencies greater than 100 kilocycles per sec. (Note that at 40 kilocycles per sec. the addition is 3 per cent.) This calibration will then hold without further adjustment between 100 and 6 000 kilocycles per sec. ($\lambda = 3\,000$ m to 50 m). In calculating the factors collected in Table 3, the inductance has been taken as 0·2 μH and the terminal capacity as 3 $\mu\mu$F. The correction for longitudinal distribution of current has been calculated from formula (5) and in this k has been taken as 2 and l as 40 cm. The current corresponding to the observed deflection is to be read from the calibration curve, and then this current is to be reduced by the amount shown in the last column of Table 3.

TABLE 3.

λ	f (kilocycles per sec.)	Correction for current distribution	Correction for terminal capacity	Net reduction factor for current
m		per cent	per cent	per cent
50	6 000	0·085	0·02	0·1
20	15 000	0·504	0·5	1·0
10	30 000	2·15	2·0	4·1
5	60 000	8·2	8·5	16·7

It will be seen that the correction factor is small, and perhaps unimportant, until the wavelength is less than 15 m, but when it is less than 10 m the factor rapidly becomes large. The non-uniformity of the complete system may possibly make the factor slightly different from that calculated by equation (5), and perhaps that analysis should be interpreted as showing that the factor will not reach 1 per cent until the wavelength is less than about 15 m, and should thereafter be used for estimating roughly the correction for much shorter waves. Possibly future experience with instruments having systems of different lengths may substantiate the applicability of equation (5) to bent systems. Also the correction for terminal capacity becomes very appreciable, and this factor cannot well be more than an approximate estimate.

If and when it becomes necessary to make accurate current measurements at wavelengths shorter than 10 m it will be necessary to use an ammeter with a shorter system; in other words the ammeter must constitute a smaller fraction of the total length of the circuit in use.

(8) FUTURE DEVELOPMENTS.

The mechanical development and testing of these two instruments has shown up certain difficulties and drawbacks. The chief mechanical difficulty is to provide supports for the moving cylinder which are heavy enough to carry the current and yet do not impose too great a control. The convenience of calibrating at, say, 500 cycles per sec. has shown that a steady-current calibration is both unnecessary and troublesome. Accordingly it is proposed to abandon the possibility of steady-current calibration, for then the instrument can be built in a way which possesses very great advantages both in construction and use. In future it is proposed to use only one cylinder, which will be placed eccentrically within the screen tube, and to measure the repulsive force between the current in the cylinder and the eddy currents induced in the screen tube.* The cylinder will then be fixed and the screen tube will become the moving member. Since all the current in the screen tube is induced in it by the field of the cylinder, no means are required for leading this current in and out. This avoids the difficulty of carrying current through thin strips, and oil immersion is no longer required: the absence of the strips will reduce the resistance of the instrument at least 10-fold. It is proposed to carry the screen tube on knife edges and the whole control will be by gravity. The whole system is uniform as the right-angle bends, necessitated by the control springs, are no longer present; and the whole length is operative in producing repulsive force. Also the relevant value of l to be substituted in equation (5) is the length of the screen tube instead of the whole length of the complete loop. Thus had the 20-ampere instrument been made with one cylinder and a moving screen tube, its effective length would have been 10 cm instead of 40 cm. It may be calculated from (5) that the correction for current distribution would then have been only 1·7 per cent at $\lambda = 2$ m. With such an instrument we could therefore ignore the effect of current distribution entirely. In addition the inductance of the system would be halved, because now the screen tube is equivalent to placing a thick copper plate in the median plane of Fig. 1 and then obliterating one cylinder. It will further be reduced because the length of the system has been shortened by the length of the control springs. Probably the inductance would be reduced from 0·2 μH to 0·07 μH. The presence of the ammeter would then make less disturbance to the circuit, and the small inductance would render the terminal capacity much less effective. Assuming an inductance of 0·07 μH and a terminal capacity of 3 $\mu\mu$F, it may be calculated that the correction for terminal capacity would be 0·6 per cent at $\lambda = 10$ m and 2·4 per cent at $\lambda = 5$ m. The calibration of such an instrument would therefore be independent of frequency between 100 cycles and 30 000 kilocycles per sec.

* *Proceedings of the Royal Society*, loc. cit., Fig. 13.

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THE PRESSURES ON THE DIAPHRAGM OF A CONDENSER TRANSMITTER IN A SIMPLE SOUND FIELD.*

By W. WEST, B.A., Associate Member.

(*Paper received 7th August*, 1929.)

SUMMARY.

It has been observed that the free air pressures of a sound wave may be increased at the diaphragm of a microphone in a ratio greater than 2 : 1 at certain frequencies, when there is concavity in the face of the microphone. A method of calculating the magnitude of this increase is applied to a condenser transmitter of the Wente type, and close agreement is found with experimental results. The calculations involve certain simplifying approximations the validity of which is discussed.

SECTION (1).

The condenser transmitter has been in use for many years in the perfected form, the design of which is due to Wente,† but although it is frequently required for the purpose of measuring sound pressures in free air its performance is usually measured in terms of a uniformly distributed pressure actuating the diaphragm. This type of instrument satisfies, to a greater extent than any other now in existence, the essential requirements to be met in a large class of acoustical measurement, as typified, for example, in the testing of loudspeakers for frequency characteristics. In order that accurate measurements of sound pressures (or intensities) may be made, it is necessary for the calibration of the transmitter to be obtained in terms similar to those required in its use. The difference that exists between the performance of a transmitter of this type when measured in terms of actual pressures on the diaphragm, and in terms of pressures in free air, has been discussed by Mr. A. J. Aldridge of the Post Office Research Section, with reference to characteristic curves obtained for a particular instrument.‡

In the present paper it is proposed to show that the characteristic for free air pressures may be deduced, with fair accuracy, from a calibration of actual pressures —when certain simplifying assumptions are made. There are two important corrections to be made; one of these is due to the augmentation of the sound pressures at the face of the transmitter as a consequence of the obstruction presented to the sound waves, and the other is required to take into account the fact that the pressure at the face of the transmitter is not necessarily the same as that on the diaphragm, which is recessed. Of these corrections the former has recently received attention,§ but the latter has to a large extent been ignored or overlooked; it is therefore the subject chiefly to be dealt with.

SECTION (2).

The calculations necessary to take into account the effect of the recess in front of the diaphragm are based on the conception of acoustical impedances and of reflection factors at the ends of a tube in which sound is sustained. The method has been applied to the measurement of acoustical impedances.† Consider a short tube the ends A and B of which are terminated by acoustical impedances having reflection factors (for velocities or pressures) of g_A and g_B respectively. The length l between the ends may either be taken as the distance between surfaces at which reflection takes place without change of phase (as at a rigid surface), or it may refer to the distance between the actual terminations of the tube. In the former case the factor g is simply a numerical ratio which may have any value between ± 1; in the latter case it is understood to be the product of such a ratio and a term $e^{-2ik\alpha}$, where $i = \sqrt{(-1)}$; $k = 2\pi$ divided by the wavelength, and α is the "end correction" of the tube (dimension length). Adopting the second notation, assume that sound is maintained by the motion at end A of an absorbing piston the velocity of which is $Ve^{i\omega t}$. The air velocity at A is the sum of the velocities at A of all the waves passing up and down the tube, namely,

$$v_A = Ve^{i\omega t} - Vg_Be^{i(\omega t - 2kl)} + Vg_Ag_Be^{i(\omega t - 2kl)} - Vg_Ag_B^2e^{i(\omega t - 4kl)} + \ldots \text{ to infinity}$$

$$v_A = \frac{Ve^{i\omega t}(1 - g_Be^{-2ikl})}{(1 - g_Ag_Be^{-2ikl})} \quad . \quad . \quad . \quad . \quad (1)$$

The pressure at end A is similarly given by

$$p_A = \rho cVe^{i\omega t} + \rho cVg_Be^{i(\omega t - 2kl)} + \rho cVg_Ag_Be^{i(\omega t - 2kl)} + \rho cVg_Ag_B^2e^{i(\omega t - 4kl)} + \ldots$$

$$p_A = \frac{\rho cVe^{i\omega t}(1 + g_Be^{-2ikl})}{(1 - g_Ag_Be^{-2ikl})} \quad . \quad . \quad . \quad (2)$$

where ρ = density of air and c = velocity of propagation of sound in air. The pressure at end B is

$$p_B = \rho cVe^{i(\omega t - kl)} + \rho cVg_Be^{i(\omega t - kl)} + \rho cVg_Ag_Be^{i(\omega t - 3kl)} + \ldots$$

$$p_B = \frac{\rho cVe^{i(\omega t - kl)}(1 + g_B)}{(1 - g_Ag_Be^{-2ikl})} \quad . \quad . \quad . \quad (3)$$

* Reprinted from *Journal I.E.E.*, 1930, vol. 68, p. 441.
† See Reference (1). ‡ *Ibid.*, (2).
§ *Ibid.*, (3) and (4).
† See Reference (5).

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The impedance per unit area at end A of the tube is

$$Z_1 = \frac{p_A}{v_A} = \frac{\rho c(1 + g_B\epsilon^{-2ikl})}{(1 - g_B\epsilon^{-2ikl})} \quad . \quad . \quad . \quad (4)$$

and the pressure ratio

$$\frac{p_A}{p_B} = \frac{1 + g_B\epsilon^{-2ikl}}{(1 + g_B)\epsilon^{-ikl}} \quad . \quad . \quad . \quad . \quad (5)$$

Equation (4) may be put into the form

$$Z_1 = \rho c\left[\frac{(1 - g_B^2) - i \,.\, 2g_B \sin 2\,kl}{1 + g_B^2 - 2g_B \cos 2kl}\right] \quad . \quad (6)$$

If reflection from end B takes place as from a substantially rigid surface, $g_B = 1$ and

$$Z_1 = -i\rho c \cot kl \quad . \quad . \quad . \quad . \quad (6a)$$

while $$p_A/p_B = \cos kl \quad . \quad . \quad . \quad . \quad (5a)$$

Now suppose that end A is open and provided with an infinite flange, then the total impedance per unit area at A is $Z = Z_1 + Z_0$, where Z_0 is the impedance per unit area due to the reaction of the air outside the tube. (The electrical analogue is therefore a simple circuit comprising two impedances in series.) If the total sound pressure applied at end A is p, then it follows that the ratio

$$\frac{p}{p_A} = \frac{Z}{Z_1} = 1 + \frac{Z_0}{Z_1}$$

and $$\frac{p}{p_B} = \frac{p}{p_A} \cdot \frac{p_A}{p_B} = \left(1 + \frac{Z_0}{Z_1}\right)\cos kl \quad . \quad . \quad (7)$$

The impedance Z_1 is given in equation (6a), and Z_0 is known to consist of a resistance component R per unit area due to radiation of sound, and a reactance component $im\omega$, where m is the added mass per unit area due to the presence of the external air. Hence

$$\frac{p}{p_B} = \cos kl - \frac{m\omega}{\rho c}\sin kl + \frac{Ri}{\rho c}\sin kl \quad . \quad (7a)$$

The average values of R and m for a circular piston of radius r have been evaluated.*

Thus $$\frac{R}{\rho c} = 1 - \frac{J_1(2kr)}{kr} \quad . \quad . \quad . \quad . \quad (8)$$

and $$\frac{m}{\rho r} = \frac{K_1(2kr)}{2(kr)^3} = \frac{\mathbf{H}_1(2kr)}{(kr)^2} \quad . \quad . \quad . \quad (9)$$

(Tables for the functions $J_1(x)$ and $\mathbf{H}_1(x)$ are available. Both are to be found in "The Theory of Bessel Functions," by G. N. Watson.)

Section (3).

Reference will now be made to measurements of the frequency characteristics of a condenser transmitter (No. 47) of the 370W type. The results of similar measurements on another instrument (No. 111) of the same type have already been published,† but they refer to the calibration of the instrument with its protecting cover in position. This cover consists of a metal cap fitting over the face of the transmitter, with a recessed wire-gauze shield covering the diaphragm. The presence of the cover somewhat increases the depth of the recess and reduces the diameter.

As the result of a recent investigation (see Reference 4) a special source of sound has been designed to enable measurements to be made at higher frequencies. This source of sound operates on the same principle as a Bell type of telephone receiver, but its diaphragm, being clamped on a diameter of 1·9 cm, has a fundamental natural frequency of about 3 600 cycles per sec. The external diameter of the instrument is limited to 4 cm. The adjustment of a small acoustical resonator in front of the diaphragm enables measurements to be made over a frequency range of 2 500 to 5 000 cycles per sec. The apparatus and methods of carrying out

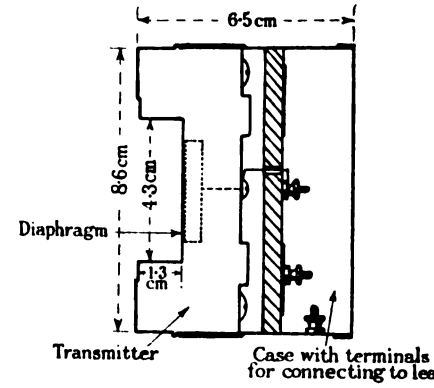

Fig. 1.—Outline of cross-section of condenser transmitter.

the measurements do not otherwise differ from those previously described.*

On the back of the condenser transmitter was mounted a short cylindrical case to provide mountings for terminating the lead to the amplifier and to act as a screen. No description of the instrument beyond that of its external geometrical shape and dimensions is necessary for the present purpose. These are sketched in Fig. 1, which shows an outline of the cross-section including the case at the back. The general form is that of a short cylinder having a cylindrical cavity in the front face.

Frequency characteristics of this instrument, obtained by means of the Rayleigh disc, are shown in curves (1) and (2) of Fig. 2, curve (1) being obtained in terms of sound pressures in free air and curve (2) in terms of actual pressures on the diaphragm. Curve (3) shows the maker's calibration (by thermophone) for the *type* and is introduced in justification of the form of extension of curve (2) beyond 3 300 cycles per sec., which was the highest frequency at which readings were taken for this condition of test. Obviously, the difference between curves (1) and (2) cannot be accounted for solely by the effect of the obstruction of the instrument to the sound waves.

* See Reference (6), par. 302.

† *Ibid.*, (2).

* See Reference (4). The transmitter was so located in the testing cabinet (Fig. 5 of this reference) that, while it faced the source of sound, secondary or reflected waves from its surface would be incident more or less at right angles to the plane of the Rayleigh disc, and would therefore not affect a measurement of the primary wave.

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Section (4).

The corrections, mentioned in Section (1) as being necessary to correlate the two curves, may be applied if certain simplifying assumptions are made, discussion of which will be deferred to Section (5). Firstly, if the existence of the cavity in the face of the transmitter is ignored, there will be at this face an increase in pressure due to obstruction of the sound field; it will be assumed that the pressure will be increased in the same ratio as would occur at the point nearest to the source of sound on a solid spherical obstacle, 10 cm in diameter. This correction factor, or ratio of actual to free air pressure, can be calculated, and it is plotted against frequency in curve (1) of Fig. 3. In making the calculations the value of c, the velocity of propagation of sound, was taken as 34 000 cm per sec. Points on the curve at frequencies above 2 500 cycles per sec. ($k \times$ radius of sphere > 2) were obtained from the table published by Ballantine (see Reference 3), the remainder chiefly from calculations supplied by Rayleigh (see Reference 6, chap. 17).

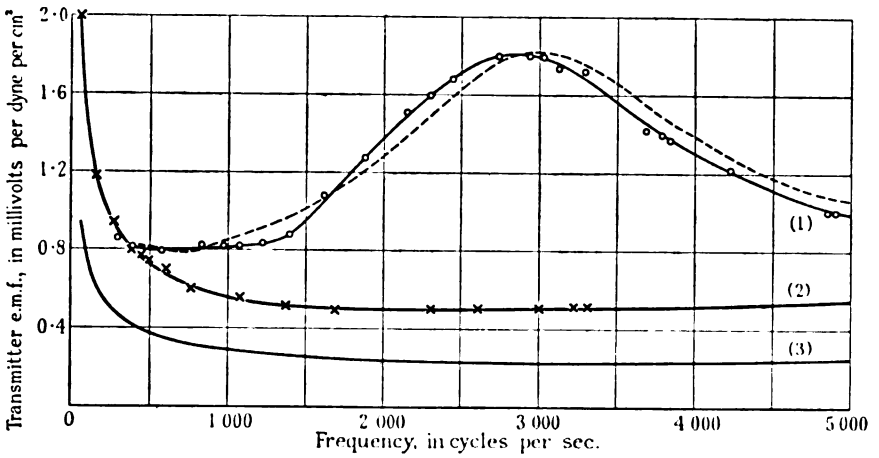

Fig. 2.—Frequency characteristics of condenser transmitters.

Curve (1).—In terms of free air pressures.
Curve (2).—In terms of pressures on the diaphragm.
Curve (3).—Maker's curve for the type (by thermophone).
Dotted curve.—Calculated curve for free air pressures (based on Curve 2).

Secondly, the ratio of the pressure p at the face of the transmitter to the pressure p_B on the diaphragm may be calculated from equation (7a), with the assumptions that the motion of the diaphragm is insignificant and that the external acoustical impedance at the aperture of the cavity is equivalent to that on (one side of) a piston enclosed by an infinite baffle. In order to simplify the calculation put $R' = R/(\rho c)$ and $m' = m/(\rho r)$, these

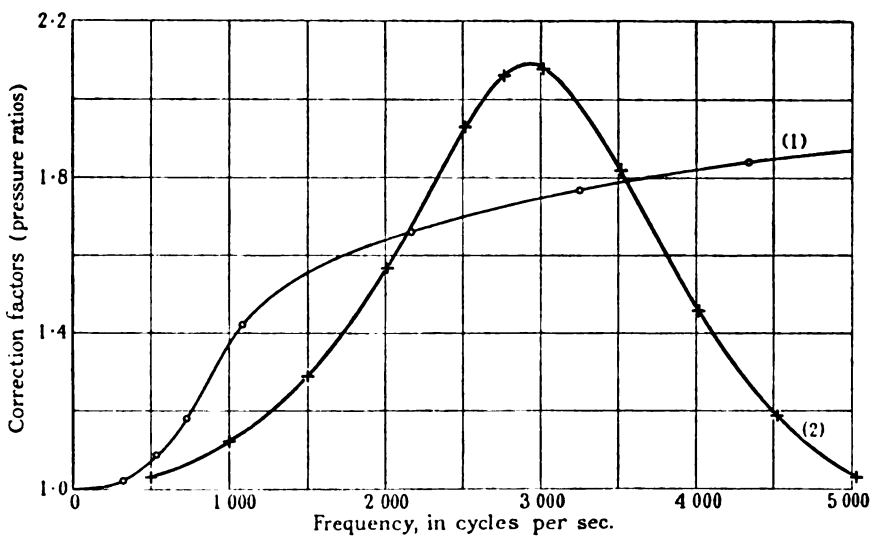

Fig. 3.—Correction factors for reducing pressures on the diaphragm into terms of free air pressures.

Curve (1).—Calculated ratio of actual to free air pressures on a sphere of 10 cm diam.
Curve (2).—Calculated ratio of pressures on diaphragm to pressures at aperture of condenser transmitter.

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values being obtained from equations (8) and (9). Then equation (7*a*) may be rewritten

$$\frac{p}{p_B} = [\cos kl - m'kr \sin kl] + i[R' \sin kl] = X + iY$$

and the numerical value of $p_B/p = [X^2 + Y^2]^{-\frac{1}{2}}$, where $X = \cos kl - m'kr \sin kl$ and $Y = R' \sin kl$.

Details of the calculations are given in the table, and the values of the correcting factor p_B/p are plotted against frequency in curve (2) of Fig. 3.

TABLE.

($l = 1{\cdot}3$ cm, $r = 2{\cdot}15$ cm)

Frequency	kr	kl	cos kl	sin kl	m'	R'	X	Y	p_B/p
500	0·2	0·12	0·993	0·120	0·840	0·020	0·973	—	1·03
1 005	0·4	0·24	0·971	0·238	0·807	0·078	0·894	—	1·12
1 505	0·6	0·36	0·936	0·352	0·770	0·170	0·773	0·060	1·29
2 010	0·8	0·48	0·887	0·462	0·714	0·288	0·623	0·133	1·57
2 515	1·0	0·60	0·825	0·565	0·647	0·423	0·460	0·239	1·93
2 765	1·1	0·66	0·790	0·613	0·610	0·495	0·379	0·304	2·06
3 015	1·2	0·72	0·752	0·659	0·571	0·567	0·301	0·374	2·08
3 520	1·4	0·84	0·667	0·745	0·493	0·708	0·153	0·528	1·82
4 020	1·6	0·96	0·573	0·819	0·414	0·837	0·031	0·685	1·46
4 525	1·8	1·08	0·471	0·882	0·338	0·947	−0·065	0·835	1·19
5 025	2·0	1·20	0·362	0·932	0·267	1·030	−0·136	0·960	1·03

The application of both of the correction factors to the calibration of the transmitter in terms of actual pressures [curve (2) of Fig. 2] results in the broken curve shown in the same figure. No ordinate on this curve differs from the corresponding ordinate on curve (1) by as much as 10 per cent.

SECTION (5).

It was first assumed that the effect of the transmitter as an obstacle in the sound field created pressures at the surface equivalent to those on a sphere 10 cm in diameter. Since such an effect has been calculated only for a spherical shape, the assumption of an " equivalent sphere " is essential. Although it is not to be expected that the true effect will take the exact form shown in curve (1) of Fig. 3, it is very probable that the differences at all points will be small, at least in the case of a short cylindrical obstacle, provided that a suitable diameter is chosen for the equivalent sphere. There is also some confirmatory experimental evidence of this (see Reference 4). In estimating the diameter of the equivalent sphere the presence of the recess may be ignored since it is given separate consideration, i.e. the cavity may be considered to be filled with solid to the front face of the transmitter. As a result of previous experience, the author has estimated that for pressures at the centre of a flat face of a short cylinder (the length of which is, say, less than the radius) the circumscribing sphere forms a useful approximation. For the transmitter the circumscribing sphere has a diameter of 10·8 cm, but the length is greater than the radius. Ballantine (see Reference 3) proposes an estimate based on equal volumes of sphere and obstacle, when the obstacle " occupies a volume of reasonable (say cubical) shape " (presumably when extreme dimensions taken through the centre do not differ too greatly). On this basis the diameter of the equivalent sphere would be 9 cm, but as this hardly exceeds the actual diameter (when taken over the slightly projecting ridge) no account would be taken of the comparatively more gradual streamline presented by the sphere, and this value would probably underestimate the effect. For these reasons the intermediate value of 10 cm was chosen for the diameter of the equivalent sphere, though either of the other values mentioned could be used without greatly affecting the result.

Secondly, the motion of the diaphragm is neglected in the assumption that $g_B = 1$, i.e. that the mechanical impedance of the diaphragm is very great as compared with the characteristic impedance of a tube (ρc per unit area). In this type of instrument the diaphragm is stretched and the fundamental natural frequency exceeds 10 000 cycles per sec. The chief constraint to motion is imposed by the air cushion at the back, for which some impedance values are quoted by Crandall.* These were calculated for an early form of condenser transmitter probably of similar dimensions and construction, and will serve to indicate the degree of approximation that has been made. For example, at 3 000 cycles per sec. the mean resistance per unit area is about 5 000 and the mean stiffness per unit area about 3×10^8 [reactance $= 3 \times 10^8 \div (2\pi \times 3\,000) = 16\,000$]. This takes no account of the stiffness of the diaphragm *per se*, or of its mass, but it will serve to assume an impedance of, say, $5\,000 - i\,10\,000$. This impedance per unit area may be related to g_B by putting $l = 0$ in equation (4), and it will be convenient to separate g_B into the form $g_B = ge^{-2ika}$, where g is simply a numerical ratio. Hence

$$\frac{\rho c(1 + ge^{-2ika})}{(1 - ge^{-2ika})} = 5\,000 - i\,10\,000$$

* See Reference (7), art. 11, Fig. 5.

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Equating real and imaginary parts,

$$\frac{\rho c(1 - g^2)}{1 + g^2 - 2g \cos 2ka} = 5\,000$$

and

$$\frac{\rho c \,.\, 2g \sin 2ka}{1 + g^2 - 2g \cos 2ka} = 10\,000$$

Since ka is very small, $\sin 2ka = 2ka$
and $\cos 2ka = 1 - \frac{1}{2}(2ka)^2$, with sufficient accuracy.

The solution of these equations gives $g = 0{\cdot}997$ and $2ka = 0{\cdot}006$, or $a = 0{\cdot}005$ cm. The sign of a is positive and the correction therefore tends to increase the length of the tube.

It may also be concluded that the performance of the transmitter cannot be appreciably affected, at the frequencies considered, by changes of the impedance imposed by the external air. Such an effect has been observed (see Reference 4) in the case of a very light diaphragm, but it applies chiefly to the position of the fundamental natural frequency, which is well beyond the frequency range dealt with and is, moreover, less liable to alteration for the greater density of the diaphragm.

The only other approximation of any consequence was that the impedance, looking outwards from the aperture of the transmitter, was assumed to be equal to the average impedance over the surface of a piston, of the same area, surrounded by an infinite baffle. This assumption is somewhat bold, but a consideration of such information as is available indicates that it is more reasonable than would appear at first sight. Consider first the resistance component, retaining for the time being the fiction of piston motion at the aperture. It is known that at very low frequencies the value of R' with a small baffle is one-half of that given by equation (8), applicable to an infinite baffle. It probably increases towards this value in much the same way as curve (1) of Fig. 2 approaches the value 2 as the frequency is raised. It can be seen from the table on page 148 that the term Y, involving this component, has little or no influence below, say, 2 000 cycles per sec., and above this frequency errors in the value of R' are probably less than 20 per cent (as deduced from curve 1 of Fig. 2). All dissipative forces other than that due to radiation have been neglected as being comparatively very small, but in so far as such do exist they will tend to reduce this error.

At low frequencies at any rate, the mass component m' is also overestimated, though not to so great an extent as the resistance component, since it plays no part in radiating sound and, being more influenced by local conditions, it probably does not require so large a baffle for the practical equivalent of one that is infinitely extended. Thus at very low frequencies the addition of an infinite baffle on the end of a tube, initially without any baffle at all, is known to increase the value of m' only in the ratio 0·82 to 0·6, or about 1·37 to 1. Since the presence of the flange forming the face of the transmitter may be expected to exert an influence even at low frequencies, it seems probable that the admitted overestimate of m' at low frequencies is rapidly reduced as the frequency is raised into the range where the correction factor becomes important. At the higher frequencies the conditions are less certain, but the term X, involving the mass, has little influence on the correction factor from about 3 500 to 5 000 cycles per sec.

In discussing the assumption of piston motion at the aperture * it may be recalled that at very low frequencies it involves an error in the "end correction," i.e. in m', in the ratio of 0·85 to 0·82 when applied to an open-ended tube with an infinite flange. This appears to be a comparatively small effect, but an element of uncertainty is introduced when the irregular distribution of impedance over the surface of the piston is considered. It is this irregularity which chiefly causes the actual motion at an aperture to depart from that of a true piston, and in assuming the impedance per unit area to be equal to the average value of that on a piston no account is taken of the form of the distribution. This is important if (the length of the tube being short) an irregular distribution of pressure is created on the diaphragm, since the effects of pressures at the centre are stronger than of those nearer the edge. Under the actual conditions, however, it is probable that variations of impedance appear as variations of velocity at different parts of the aperture rather than as variations of pressure. Warren † has drawn attention to the differences in impedance at the centre and at the edge of a circular piston. He shows that the impedance at the centre is greater than that at the edge, and observed increases in amplitude or velocity towards the edge of an aperture may thus be accounted for.

Section (6).

The close agreement reached between the calculated and the experimental curves has the double value of inspiring confidence in the methods of calculating the corrections, and of providing an additional check on the accuracy of the Rayleigh disc methods of calibration, there being (as far as the author is aware) no direct alternative method at present available for effecting a calibration in terms of free air pressures.

In a more recent form of Wente condenser transmitter the recess has been reduced to 0·85 cm in depth, the radius remaining the same. That the effect of the recess is still appreciable is shown by a calculation at 3 500 cycles per sec., at which frequency the correction factor for the recess has a value of 1·63, which is probably nearly the maximum value attained.

The calibration and use of a transmitter in terms of free air pressures require that sound shall emanate from a single source; this necessitates the absence of reflecting surfaces, that is the carrying out of experiments in open air, or the acoustical equivalent. The latter alternative is available by covering all the solid surfaces in a testing cabinet with a sufficient quantity of a material suitable for accepting and absorbing sound.

The value to acoustical engineering practice of the conception of acoustical impedances is generally recognized; it would be increased by a more definite knowledge of the actual magnitudes of the impedances in some

* At the lowest frequency calculated (500 cycles per sec.) the assumption of piston motion results in a value (0·84) of m' in excess of that calculated by Rayleigh for an open-ended tube with an infinite flange, namely 0·82, but at this frequency the correction is, in any case, very small.

† See Reference (8).

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common cases (e.g. of apertures surrounded by finite flanges) where direct calculation has failed. Experimental methods are available which would serve to furnish data of this kind.

The measurements which have been recorded were carried out at the Post Office Engineering Research Station and are published by permission of the Engineer-in-Chief of the Post Office.

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(5) W. WEST: "Measurements of the Acoustical Impedances of Human Ears," *Post Office Electrical Engineers' Journal*, 1929, vol. 21, p. 293.

(6) LORD RAYLEIGH: "Theory of Sound," vol. 2 (Macmillan, 1896).

(7) I. B. CRANDALL: "Theory of Vibrating Systems and Sound" (Macmillan, 1927).

(8) A. G. WARREN: "A Note on the Acoustic Pressure and Velocity Relations on a Circular Disc and in a Circular Orifice," *Proceedings of the Physical Society of London*, 1928, vol. 40, p. 296.

MEETINGS OF THE WIRELESS SECTION.

80TH MEETING OF THE WIRELESS SECTION, 15TH JANUARY, 1930.

Captain C. E. Kennedy-Purvis, R.N., Chairman of the Section, took the chair at 6 p.m.

On the motion of the Chairman, votes of condolence were passed on the death of Dr. S. Z. de Ferranti, F.R.S. (a Past President and Honorary Member), and Admiral Sir Henry B. Jackson, R.N., G.C.B., K.C.V.O., F.R.S., the members standing in silence.

The minutes of the meeting of the Wireless Section held on the 4th December, 1929, were taken as read and were confirmed and signed.

A paper by Mr. H. A. Thomas, M.Sc., Associate Member, entitled "A Method of Measuring the Overall Performance of Radio Receivers" (see page 81), was read and discussed.

The meeting terminated at 8 p.m. with a vote of thanks to the author, which was moved by the Chairman and carried with acclamation.

81ST MEETING OF THE WIRELESS SECTION, 5TH FEBRUARY, 1930.

[*Held at the City and Guilds (Engineering) College.*]

Captain C. E. Kennedy-Purvis, R.N., Chairman of the Section, took the chair at 6 p.m.

The minutes of the meeting of the 15th January, 1930, were taken as read and were confirmed and signed.

The following papers were read:—

"An Ammeter for Very High Frequencies." By Prof. C. L. Fortescue, O.B.E., M.A., Member, and Mr. L. A. Moxon (see page 102).

"A Method of Measuring Mechanical Impedance." By Prof. E. Mallett, D.Sc.(Eng.), Member, and Mr. R. C. G. Williams, B.Sc.(Eng.), Graduate (see page 106).

"Some Notes on the Design of a Gramophone Pick-up." By Mr. G. W. Sutton, B.Sc. (see page 112).

"Frequency Stabilization of Valve Oscillators." By Prof. E. Mallett, D.Sc., Member (see page 124).

"An Instrument for Projecting and Recording the Response Curves of Electrical Circuits." By Prof. C. L. Fortescue, O.B.E., M.A., Member, and Mr. F. Ralph (see page 129).

An adjournment was then made to the laboratories of the College, where the apparatus described in the above papers was demonstrated and informally discussed.

At the dinner held at the Imperial College Union after the meeting the Chairman proposed a vote of thanks to the Delegacy of the College, to which Mr. W. M. Mordey (one of the Governors of the College) and Prof. C. L. Fortescue responded.

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SOME DEVELOPMENTS OF THE PIEZO-ELECTRIC CRYSTAL AS A FREQUENCY STANDARD.*

By H. J. Lucas, Associate Member.

(*Paper first received 16th December, 1929, and in final form 31st January, 1930; read before the* Wireless Section *5th March, 1930.*)

Summary.

Part (1) deals with some observed errors in quartz piezo-electric resonators employed as frequency standards. It is shown that these are due to (*a*) frictional loading, and (*b*) losses arising from atmospheric humidity. The methods adopted to eliminate these are discussed, and the development of calibration apparatus employing the resonators as reference standards is dealt with.

Part (2) is in the main concerned with the use of the improved quartz crystal as a control element in a valve-maintained source of oscillations. The frequency-stability characteristic of such a system is investigated, the errors are found to be largely due to the valve system, and methods designed to eliminate these are discussed.

Finally a complete calibration equipment consisting of a multivibrator system with a range of 1–6 000 kilocycles per sec. in steps of 1 kilocycle per sec., centrally controlled by a quartz crystal, is shown.

Introduction.

Up to and including the War period the author was interested in the maintenance of a sub-standard wavemeter for the calibration of wavemeters in bulk. An accuracy of 1 per cent was found sufficient for all needs at that time. In the years immediately following the War accuracies up to 1 part in 1 000 were called for, while at present the sub-standard has to be correct to within 1 part in 10 000.

The first phase calls for no comment. The requirements of the second were met by the use of a wavemeter employing an oscillating triode. The accuracy of this was maintained by reference to a calibrated low-decrement closed oscillatory circuit. The latter has since become known as an " absorption wavemeter " and the reference was carried out by means of the phenomenon known as the " Ziehen " effect.

The third phase is met by similar methods, but the reference standards are in this case quartz resonators. The first part of the paper deals with the development and applications of these reference standards.

Part (1).

Summary of the Relevant Points in the Modern Theory of Coupled Circuits.

In practice it is found that the application of the " Ziehen " effect in the second phase and the various applications of quartz oscillators and resonators in the third phase are governed by the same fundamental principles. Nearly the whole of the practical applications of quartz crystals conform to the laws of coupled circuits. With suitable adjustment of the circuit these, in common with magneto-striction devices and valve-maintained tuning forks, exhibit the characteristic frequency-discontinuities of a system consisting of a low-decrement circuit tightly coupled to a source of free oscillations.

In certain instances the control member is used to obtain retroaction over a narrow frequency band and simultaneously to stabilize the resulting oscillations, but in no sense does it act as a generator.

The equivalent coupled-circuit conditions for the quartz crystal have been deduced by D. W. Dye † and other writers. The following relevant points in the theory of coupled circuits have been taken from papers by J. S. Townsend ‡ and H. G. Möller.§ A coupled circuit with a fixed primary impedance and a secondary capable of being tuned to resonance with the frequency of the applied e.m.f. is fed by a valve or other source of free oscillations. Provided that the coupling is greater than the value defined by Möller as the " critical " value, continuous variation of frequency with secondary tuning cannot be obtained. Abrupt frequency discontinuities appear on either side of the resonant points —Möller's " Ziehen " effect. Fig. 1 illustrates this point.

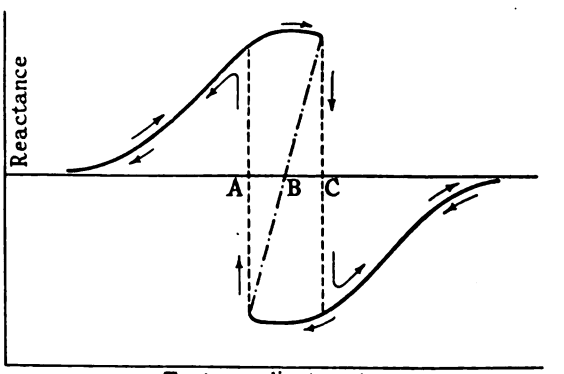

Fig. 1.

It will be noted that the portions of the operating curves bounding the hysteresis area show a very marked frequency stability, a comparatively large change of tuning producing but a small frequency change. A mathematical analysis of the simple coupled-circuit as described shows that there are three possible modes of oscillation, f_1, f_0, f_2, where f_0 is the frequency of the applied e.m.f. with the secondary removed. An examina-

* Reprinted from *Journal I.E.E.*, 1930, vol. 68, p. 855.

† " The Piezo-Electric Quartz Resonator and its Equivalent Electrical Circuit," *Proceedings of the Physical Society of London*, 1926, vol. 38, p. 399.

‡ " Oscillations obtained by Coupling a Secondary Circuit with a Continuous Wave Valve Oscillator," *Radio Review*, 1920, vol. 1, p. 369.

§ " Zur Theorie des Ziehens," *Jahrbuch der Drahtlosen Telegraphie*, 1920, vol. 16, p. 402.

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tion of Fig. 1 shows that in the vicinity of resonance (B) the mathematically possible oscillation, f_0, cannot in practice be obtained, whereas f_1 and f_2 can be obtained at will. Möller gives a reasoned case as to why these conditions obtain.

The deductions from these observations are as follows:—

(1) All mechanical vibrator-valve combinations which obey the laws of coupled circuits produce frequency-stabilized oscillations.

(2) Such combinations have two possible modes of oscillation. In the case of quartz either may be obtained by connecting the crystal to (*a*) grid-filament, or (*b*) grid-anode.

(3) The combination cannot oscillate at the natural free period of the mechanical vibrator.

The complete argument may not be strictly true in all cases of quartz-valve combination, as the frequency break, while invariably present, is displaced, and f_1 and f_2 are higher in frequency than f_0. A very close approximation to f_0 may be obtained by the Cady crevasse method, but in this case the crystal is not used as a " control " member.

A further point may be made at this juncture. The cases referred to above are in general for free oscillations, but the outputs from such combinations behave as forced oscillations, and they may be used to control other free oscillations of approximately the same periodicity and certain far-removed periodicities, as will be discussed later.

First Application of the Quartz Resonator.

In view of the lack of uniformity in the nomenclature used in the literature on the subject of quartz crystals, in accordance with Dye's practice the following conventions have been adopted:—

(1) Longitudinal vibrations—the mode of oscillation employing the longest dimension of the crystal. The cut for this may vary, but for the purposes of definition a cut normal to any two parallel faces of the natural crystal may be accepted.

(2) Transverse vibrations. This mode of oscillation employs the thickness of the crystal, i.e. it is at right angles to the longitudinal mode.

(3) Resonator—a crystal used non-retroactively. In general the longitudinal mode is employed.

(4) Oscillator—a crystal used retroactively or to control a retroactive circuit. The transverse mode is usually the more effective in this connection.

Practically the whole of the investigation which follows is concerned with the longitudinal mode of vibration. Considered purely from the precision standpoint, the longitudinal mode presents great advantages over the transverse. As an instance, the temperature coefficient of the former is only about one-twentieth of that of the latter. This important fact was first pointed out by Dye.

The introduction of the quartz resonator by W. G. Cady * was welcomed by the author with considerable relief. It was appreciated that the closed-circuit reference standards then in use were not to be relied upon for the accuracies demanded, and quartz appeared to offer a completely reliable alternative. Its inherent nature suggested absolute permanence and stability, its natural period could be determined with a high degree of accuracy and its reactions on the oscillating circuit were sharp and

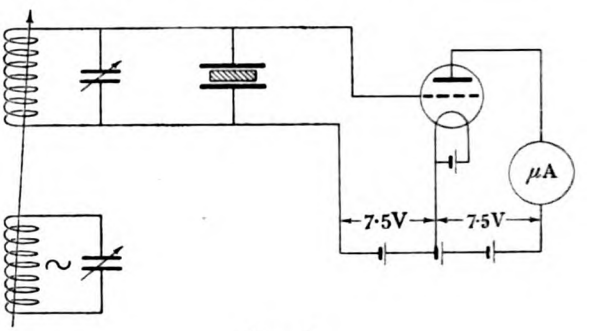

Fig. 2.

positive. An ideal solution appeared to have been arrived at, and for a time it was accepted without question. However, certain variations, apparently of the sub-standard wavemeter, were observed and these

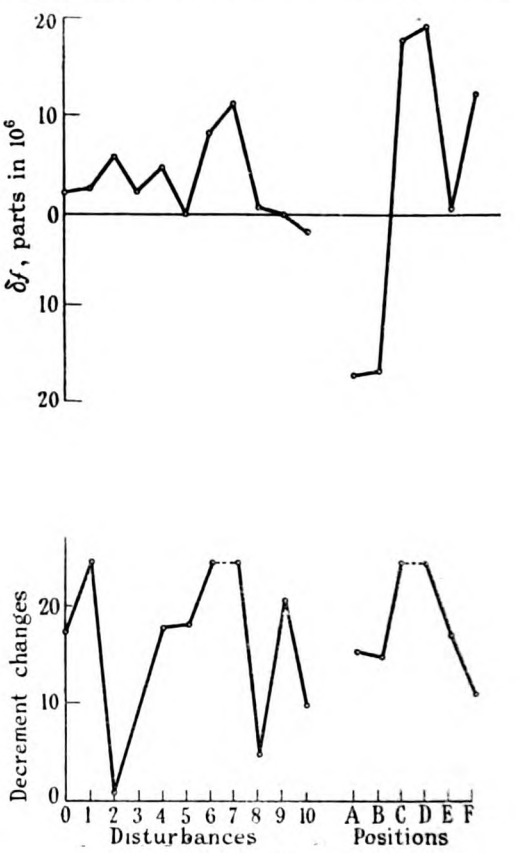

Fig. 3.

(of the order of 1 part in 3 000) could not be definitely traced to the wavemeter. It was decided, therefore, that an investigation into the behaviour of the quartz resonator was necessary.

Reference to a multivibrator confirmed the opinion

* "The Piezo-Electric Resonator," *Proceedings of the Institute of Radio Engineers*, 1922, vol. 10, p. 83.

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that the resonator or its mount was responsible for the variations. Suspicion naturally fell on the mount and a full investigation was undertaken to determine the cause of the variations.

The method of observation was that of the Cady crevasse, the only difference from the then standard practice being the use of a heavily biased valve voltmeter instead of the more usual thermo-junction and galvanometer. The circuit is shown in Fig. 2. This enabled less power to be used in the oscillator and gave a more sensitive record of the minute variations of potential in the operating region at the bottom of the crevasse. Fig. 3 shows a typical series of observations. The logarithmic-decrement variations shown are only relative, as they were disclosed partly by changes in the depth of

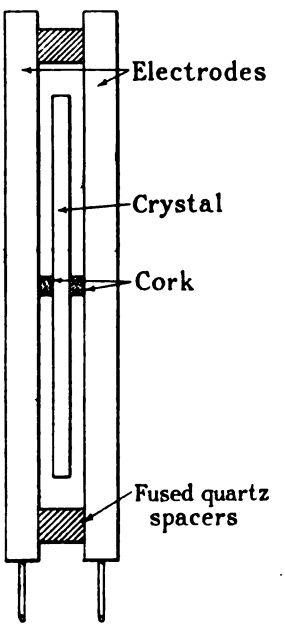

FIG. 4.

the crevasse and partly by the sharpness of what is, in effect, the crest of the resonance curve of the resonator. It was evident that within the limits of 1 part in 3 000 the crystal was very unstable. In fact the slightest disturbance in the laboratory, such as closing a door, altered the readings. It was known that the frequency variations could, in part, be accounted for by changes in the small air-gap, but this would not account for the decrement changes.

To investigate this point the resonator was supported centrally in a large air-gap (1 mm each side). The variations were found to be worse, and the decrement was also increased. This gave the clue to the trouble, as the supports for centring the crystal were obviously adding to the existing sources of error. These were finally traced to the effect of frictional loading between the crystal and its supports, which was altering both the period and the damping factor of the resonator.

The solution was simple. The resonator was firmly gripped at a nodal point of mechanical displacement for the fundamental longitudinal mode (in the centre of its length), so as to leave it free to vibrate without friction. Fig. 4 is a plan view of a resonator mounted in accordance with this method. The whole of the short-period variations shown in Fig. 3 had now vanished and the resonator could be placed in any position or could be jarred without alteration of frequency or decrement, and readings could be reproduced with a high degree of accuracy. For the first time a really consistent plot of the Cady crevasse could be obtained, enabling the numerical value of the logarithmic decrement to be computed.

This specially-mounted resonator was kept under observation for some time and long-period variations of frequency, of the order of 1 part in 10 000, as well as decrement changes were noted. These were eventually associated with changes in atmospheric humidity and

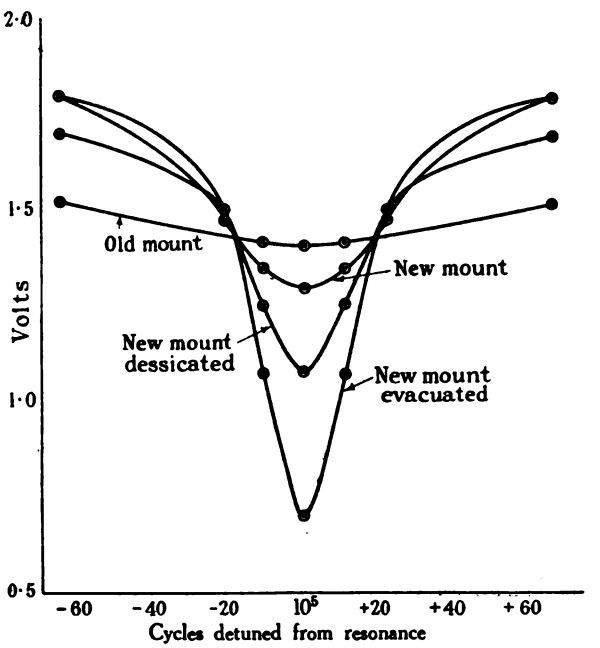

FIG. 5.

the relation between the cause and its effect was soon confirmed by tests in:—

(1) a saturated atmosphere (when the piezo-electric effect disappeared), and
(2) a sealed test-tube, with a desiccating agent.

In the latter case definite stability was obtained, after a period of slow drift during which the moisture was absorbed, together with an improvement in the decrement. An increase in the frequency of approximately 1 part in 10 000 was also noted; this point will be further discussed in Part (2). The frequency change may be partly due to the removal of the loading effect of the film of moisture on the surface of the crystal, while the improvement in decrement may result from the elimination of the hysteresis loss in this film. Dye has shown that the equivalent capacity of a similar resonator is of the order of $0 \cdot 08\ \mu\mu F$. As the logarithmic decrement is a function of the capacity as well as of the resistance, it will be realized that a small loss in such a minute capacity will have a marked effect upon the decrement. Dye comments upon the probable effect of humidity on

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frequency and has also adopted a form of nodal-support mounting.*

The obvious step of evacuating the enclosure which contained the resonator was now taken, and the anticipated further improvement in the decrement due to the removal of the air-loading was obtained. It is doubtful whether evacuation or even desiccation without the mounting precautions would be of much value. The value of the order of 1 part in 5 000 which is not at all constant. During the investigations described in the foregoing it was found possible to design and build valve-maintained oscillators which would remain constant in frequency for reasonable periods to within 1 part in 100 000. A typical time/frequency curve for such an oscillator is shown in Fig. 6.

It was found that the principal causes of the variations

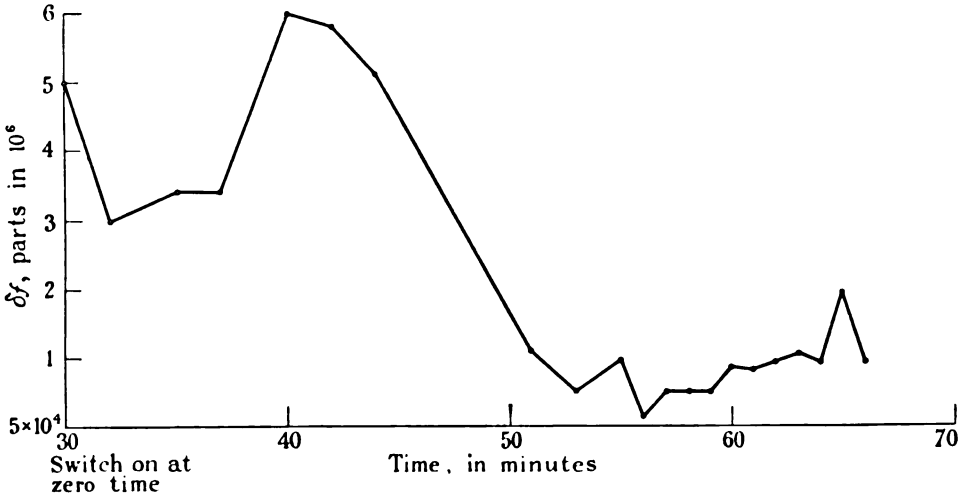

FIG. 6.

frequency variations and decrement depreciation due to frictional loading would to a large extent mask the improvement due to the removal of air loading. Fig. 9 shows the improvements in decrement as indicated by the depth and sharpness of the crevasse, produced as a result of the successive stages of development. The frequencies have been reduced to a common value for

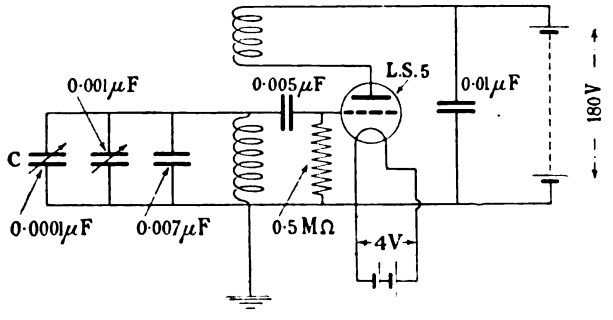

FIG. 7.

convenience—in practice these differ to the extent of 1 part in 5 000 for the extreme cases.

The " figure of merit " of a resonator or similar system is the degree of accuracy to which the true resonant frequency can be determined. This in turn is related to the logarithmic decrement. In the case of the mounted and evacuated crystal, Fig. 10, an open-scale plot of the bottom of the crevasse, shows that this determination can be made to an accuracy of 1 part in 200 000, whilst the normally-mounted crystal gives a

* Pp. 417 and 447, *ibid.*

which are usually attributed to this system were changes in the capacity of the variable condenser for any given setting, together with alterations in the valve characteristic. The inconsistencies of the variable condenser could be reduced by careful design of that component, and the residual effect could be diminished to negligible proportions by the use of a fixed shunting-capacity, which should be as large as is consistent with the provision of the necessary tuning-range. Variations of valve characteristic and of anode and filament-supply potentials were all shown as changes in the equivalent capacity of the valve. As this capacity was in parallel with the tuned circuit its effect was reduced by the provision of a shunt capacity in the latter. The circuit and values used are shown in Fig. 7. This circuit could be checked back to the quartz resonator when necessary by using it as the variable source in the Cady crevasse method, and it was found desirable to distort its output to produce the requisite number of harmonics for use in calibrating other oscillators. Further development of this method produced the equipment of which the circuit is shown in Fig. 8.

The 10-kilocycle oscillator was built on the principles set out above, while the power amplifier followed normal practice. The 10th harmonic of the distorting circuit was selected by a tuned circuit and referred to the 100-kilocycle quartz resonator by the Cady crevasse method. The adjustment was carried out by means of the fine tuning condenser C (Fig. 7), while the output from the coupling coil L gave a full range of harmonics from 10 to 2 000 kilocycles per sec. in steps of 10 kilocycles. As the adjustment of the oscillator frequency was carried out by referring the distorted output to the

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resonator, any errors due to coupling were allowed for. The rectifying valve was found to be the most convenient method of producing the requisite number of harmonics.

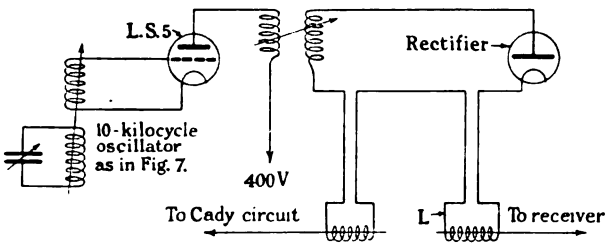

FIG. 8.

The fine-tuning condenser was calibrated in terms of the number of cycles per second by which it departed from the resonator frequency.

At the time when this equipment was developed, the B.B.C. stations were all, with the exception of 5 XX, working on a frequency which was a multiple of 10 kilocycles. Hence each station was nominally on a harmonic of the 10-kilocycle oscillator and any departure could be noted by the difference between the 10-kilocycle adjustment of the fine-tuning condenser and that obtained by referring the appropriate harmonic to the frequency of the station's carrier wave by the double-beat method; this difference was multiplied by the harmonic order of the station relative to the oscillator. Some interesting observations were made on the accuracies and stability of these stations and it was particularly noticed that 5 GB, whilst seriously departing from its nominal frequency, showed remarkable constancy. This was subsequently confirmed by Captain Eckersley's paper.* The result of a series of observations on 2 LO is shown in Fig. 9, in which the curve commences at the instant at which the station begins to radiate.

This equipment was found of very great use in the calibration of sub-standard wavemeters to the required degree of accuracy. It was also used for the measurement of inductance and capacity at various high frequencies. Further, it was and still is considered that the ideal use of a frequency standard is as a means of adjusting the frequency of a separate oscillating system and not as a "drive" or "control." The analogy to the pendulum is the employment of a free undriven pendulum as a check on a driven system, by some means which does not cause the free pendulum to deviate from its true period. It is believed that modern methods of high-precision horological determination have developed in this direction.

Briefly, the advantages of a system employing a resonator as a reference standard are:—

(1) The resonator is not doing any work—it is fed with just sufficient energy to enable the determination to be made, and the load applied by the indicating device (valve voltmeter) is practically one of potential.

(2) Provided that the "Cady crevasse" closed circuit is accurately tuned, the true period of the crystal-mount assemblage is obtained.

(3) Any variation in the electrical circuit is immediately made manifest and is under complete control.

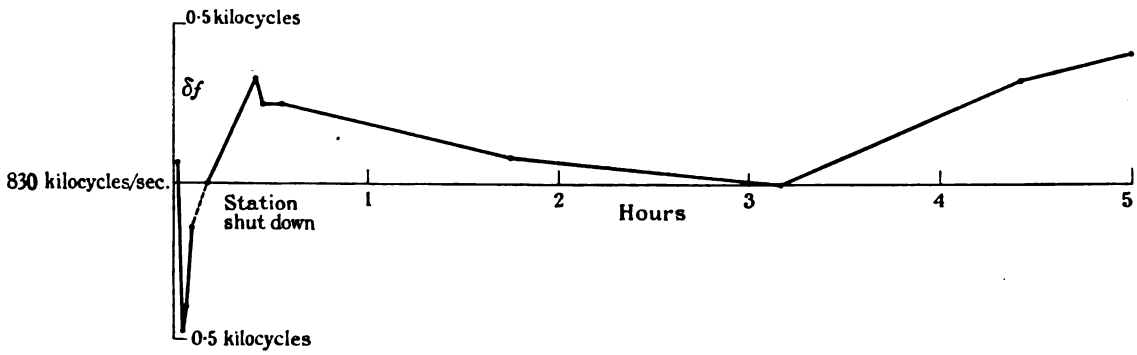

FIG. 9.

The principal disadvantage of the system is a slight uncertainty in the value of the true resonant-period due to the approximately straight line which exists at the bottom of the crevasse and which at present limits the accuracy of adjustment to about 1 part in 200 000 (Fig. 10). It is considered, however, that this could be

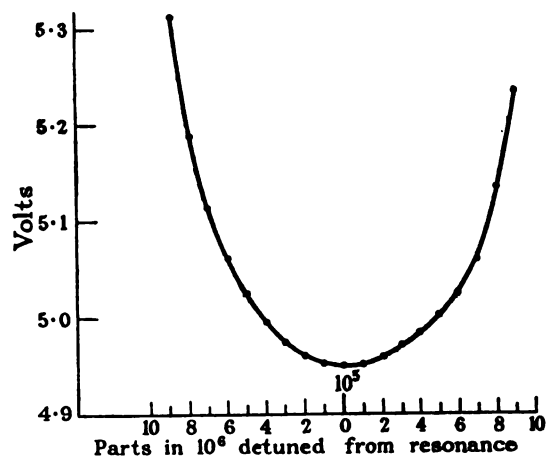

FIG. 10.

* "The Design and Distribution of Wireless Broadcasting Stations for a National Service," *Journal I.E.E.*, 1928, vol. 66, p. 501, also *Proceedings of the Wireless Section*, 1928, vol. 3, p. 108.

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greatly improved by further research, one suggestion being the employment of a high-frequency amplifier between the resonator circuit and the valve voltmeters. It has been found necessary to postpone these experiments owing to the need for the developments discussed in Part (2).

The fork-driven multivibrator had proved to be such a useful piece of apparatus that the possibility of applying the quartz standard to a similar scheme presented an irresistible attraction. Also a source of precise forced oscillations was required for general investigational work. It appeared that a great deal of research work would be needed to enable the resonator system to be used with the accuracy which, it was felt, would be called for in the near future. A wide-range low-frequency oscillator was also necessary for interpolation between the 10-kilocycle harmonics.

Further consideration of the resonator was therefore postponed and the work which is discussed below was undertaken, with the intention of returning to the reference standard in the near future.

Part (2).

Desiderata of the Drive Circuit.

(1) Experience has shown that any attempt to introduce retroaction produces a tendency to frequency instability, therefore the circuit should be capable of oscillating with no retroaction with the exception of that inherent in it.

(2) The controlling member should have a low temperature-coefficient.

(3) The coefficient of coupling between the mechanical component and the valve system should be of as low a value as possible.

In applying these principles, (1) is met by a tuned anode circuit with the crystal connected between the grid and the filament. Oscillations are more readily produced by the grid-anode connection but they are feebler and less stable than those produced by the alternative connection. (2) is ensured by the use of bars employing the longitudinal mode of vibration. This gives a temperature coefficient of the order of 2 or 3 parts in 10^6 per degree C. The transverse modes give values of the order of 40 to 70 parts in 10^6. The reason why these latter modes are frequently used is that serviceable oscillators are more easily produced from them than from the longitudinal mode. (3) is met by the use of air-gaps of the order of 1 mm between the crystal and its electrodes. In any case the electrostatic coupling of the crystal is much less than that of magneto-striction, fork, and other such electromagnetically-driven devices.

Development of the Drive Circuit.

Preliminary work showed that very good samples of this type of crystal would produce oscillations under the stipulated conditions provided that no air-gap was used, but that they failed to function when nodally mounted with an air-gap. Utilizing the experience gained on the resonators, some of these crystals were mounted in a desiccating chamber and the best of them were found to oscillate. One of these was used in a sealed tube for the preliminary development work, and the only drawbacks noted in its behaviour were a failure to oscillate below 10° C. and the fact that it would not operate with a grid-leak and had perforce to be used under free grid conditions. At a later stage the evacuated type was employed and these limitations disappeared. To eliminate the possibility of frequency variation due to grid contact-potential, bombarded, non-" gettered " valves

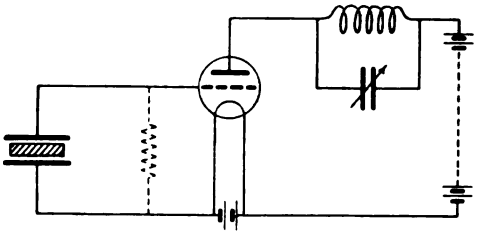

Fig. 11.

of the L.S.5 class have been used in all the later stages of the work. The drive circuit is given in Fig. 11 and presents no novel features.

Frequency Stability of the Drive System.

The reference standard used for the majority of the work is a valve-maintained tuning fork. An N.P.L. Certificate gives its period, 1 kilocycle per sec., to the seventh figure, the last being " dropped," and a zero temperature-coefficient over the working range of temperature. It is considered that the former can be relied upon to 3 parts in 10^6, and such accuracy met the

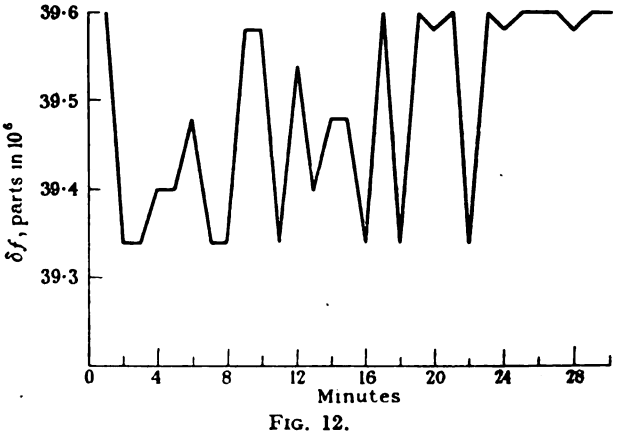

Fig. 12.

needs of the preliminary investigations. The method adopted for referring the drive output to the reference standard was as follows:—

The fork was used to drive a 1-kilocycle multivibrator circuit through a biased coupling-valve. The 50th harmonic of this was beat against the drive circuit and the resultant beats were counted over a definite period of time.

Determinations of the period of such systems as these are usually effected by reference to a time standard by means of some form of synchronous motor. The method has the advantage that the comparison may be made with a fundamental time standard, but it suffers from

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two disabilities:—(*a*) the intervention of the synchronous motor has the effect of masking any short-period variations which may be present, and (*b*) a run covering a number of hours, and in some cases several days, is necessary to obtain a determination to within 1 part in 10^6. The method outlined above enables a determination to 1 part in 10^6 to be obtained in time periods of the order of 1 minute, for in this time (3×10^6) cycles are completed and any difference between the crystal circuit and the 50th harmonic of the fork appears as beats. Extended experience has shown that observations to 1 in (3×10^6) of beats of the order 3 per second are easily possible by taking counts over periods of 1 minute. To produce an error of this order requires an uncertainty in the chronometer reading of 0·4 sec. in 1 minute. In general, 200 beats are timed.

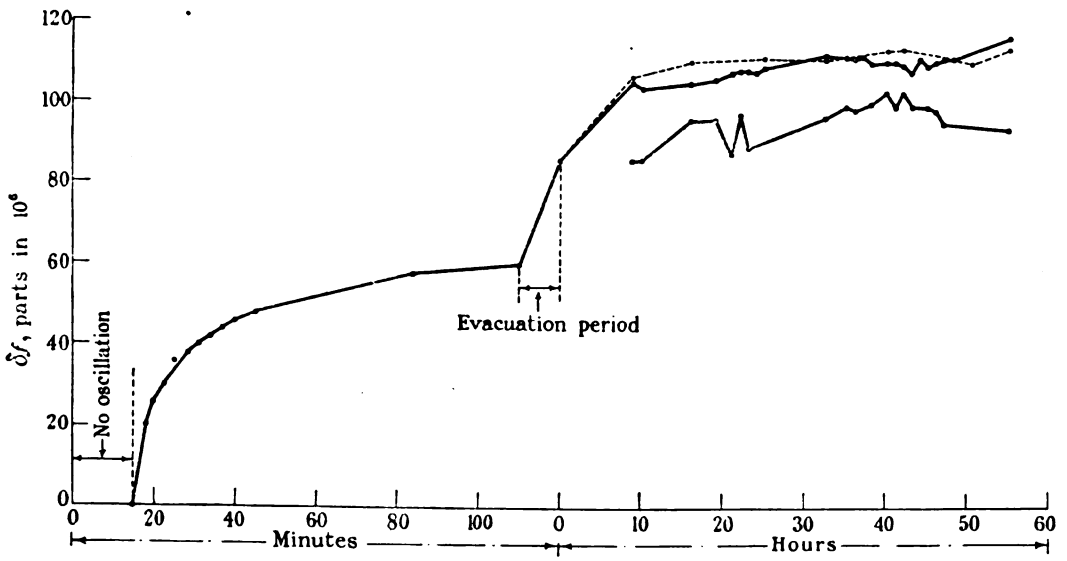

FIG. 13.

This method also permits observation of short-period variation, and the author believes that these are the first records of this nature. The crystal oscillator is adjusted by means of the anode condenser and other controllable variables so as to obtain a round number of beats in 1 minute, say 100. The count is carried on continuously and it is only necessary to note and record the variation of the chronometer seconds-hand from the 60-second mark on the completion of each hundred beats. Here again, possible errors introduced by slight inaccuracies of the chronometer reading are much less than the uncertainties due to the reference standard. An instance of an observation of this character is given in Fig. 12.

The method suffers from two defects:—(*a*) the reference is not direct to a fundamental time-standard, and (*b*) it is only applicable where the system under observation is sufficiently close to the standard to produce beats of the order 4 per second, or less. This latter condition has in all cases been easily met in these investigations.

The first observations were carried out on the non-evacuated crystal, using a free grid circuit. It was promptly found that the frequency was varying in an erratic manner and no consistent readings could be obtained. This condition was explained by reference to some work * carried out by the author's colleagues on the operation of valves under free grid conditions. The remedy consisted in bonding and earthing all free metal parts in the vicinity of the grid circuit. A permanent frequency-change of about 25 parts in 10^6 was noted and the system was quite stable.

This system was in use for some weeks and no frequency change greater than 5 parts in 10^6 was noted. Indeed, when a biased-grid system was adopted later it was found to be a most difficult matter to obtain an accuracy of this order. The reason is not far to seek. The free-grid system operates in such a manner that no grid current is flowing and there is a very close relation between grid current and frequency change. It is

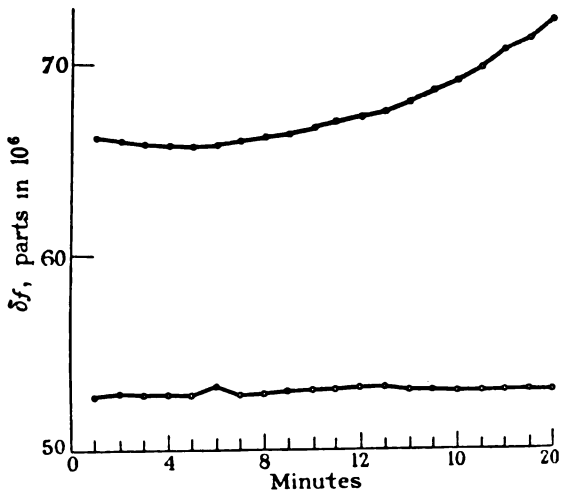

FIG. 14.

* P. L. WOSTEAR and R. H. BILLINGSLEY: "Improvements Relating to Thermionic Valve Circuits," British Patent No. 12880—1926.

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considered that such a system, properly screened, would be suitable for further investigation. It was felt, however, that the conventional biased-grid would prove more generally acceptable. In the meantime the system had been sufficiently improved to operate with grid bias.

The lower curve of Fig. 13 and the upper curve of Fig. 14 show some observations on this biased system. Fig. 13 shows a number of intermittent readings taken over 50 hours. The general upward drift is largely accounted for by temperature. Rapid fluctuations, however, are present. Fig. 14, upper curve, shows the results of a 20-minute run under continuous observation, and one of the larger excursions from the mean value shown in Fig. 13 is evidently taking place. The magnitude of these changes caused surprise and the system seemed definitely inferior to the free-grid system in previous use. As at the time no plausible reason could be offered for these fluctuations, an investigation of all the variables under control was undertaken.

Frequency-Changes Due to the Controllable Variables.

(*a*) *Anode Variable-Condenser.*—Fig. 15, top curve, gives a typical example; the other two curves show grid and anode currents. This demonstrates that while the

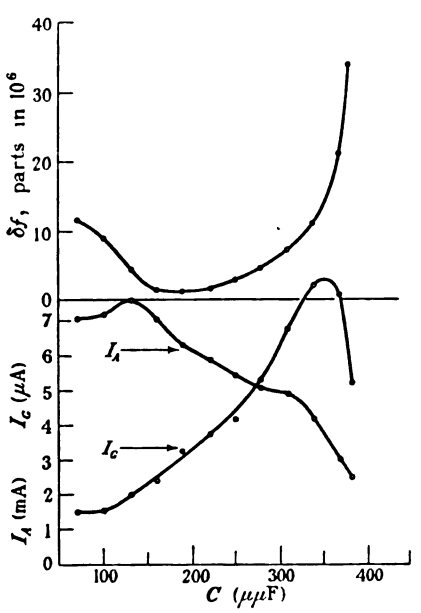

Fig. 15.

frequency is by no means independent of the condenser setting, it is practically independent of any small fortuitous capacity-variation for a given setting. An appreciable range of capacity gives $df/dC = 0$, but unfortunately this coincides with minimum output conditions. Maximum output occurs, just before oscillations cease, at 400 $\mu\mu$F. The form of the curve obtained agrees very well with that which might be deduced from the theory outlined in Part (1), this curve being a part of one of the boundary curves in the region of instability in Fig. 1. In this instance, if a similar crystal were also connected to the grid and anode, on the extinction of the oscillation f_1 at the point of maximum capacity a new frequency f_2 would develop. These points, however, have no bearing on the question of frequency stability.

(*b*) *Anode-Inductance Variation.*—Table 1 shows that the system is independent of any possible normal variation of this nominally fixed component.

Table 1.

Inductance	Time for 100 beats	Variation of frequency from 5×10^4 cycles per sec.
μH	secs.	parts in 10^6
27 000	33·4	59·9
12 400	30·0	66·6
35 000	33·8	59·2

(*c*) *Anode-Potential Variation.*—In Fig. 16 a working range of 2·1 to 1·8 volts per cell has been assumed as the possible maximum voltage-variation. This gives a variation of 105 to 90 volts for a nominal 100-volt accumulator battery, and the frequency variation for this is 7 parts in 10^7. Fig. 17 is a similar curve for the standard fork which shows a frequency variation of 1 in 10^5, a ratio of 1 : 14 in favour of the crystal.

(*d*) *Filament-Potential Variation.*—Figs. 18 and 19 show the variation of frequency with E_f for the crystal and fork respectively. In this case the comparison is

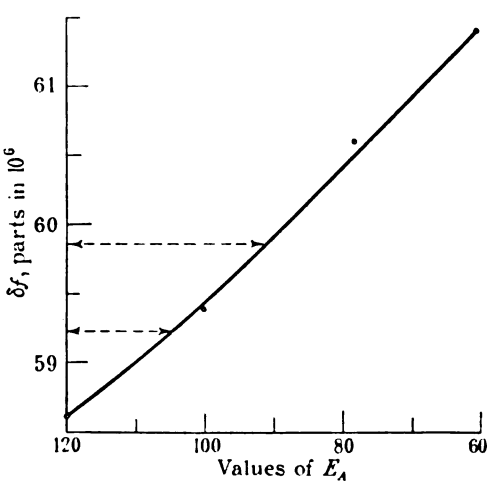

Fig. 16.

even more in favour of the crystal, the ratio being 1 : 70. Over the working range of voltage the frequency change for the quartz is 1 part in 10^7, and for all practical purposes this is negligible.

(*e*) *Grid-Potential Variation.*—This was effected by changing the value of the grid-leak, or by connecting a cell in series with it. Table 2 gives the frequency values obtained. In practice, such changes in the grid conditions are unlikely to develop.

An analysis of these results, while illuminating in themselves, gave no clear guidance as to the cause of the variation shown in the lower curve of Fig. 13.

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General observation, however, gave rise to the following considerations:—

Whilst no definite proof was forthcoming that the changes were not due to the crystal or its mount, it was felt, particularly as the resonators which had been under

TABLE 2.

Conditions	Time for 100 beats	Variation of frequency from 5×10^4 cycles per sec.
	secs.	parts in 10^6
Grid-leak A = 3MΩ; no bias ..	33·4	59·9
Grid-leak A = 3MΩ; − 1·5 V bias	32·5	61·6
Grid-leak A = 3MΩ; + 1·5 V bias	32·2	62·1
Grid-leak B = 2MΩ; + 1·5 V bias	31·0	64·5
Grid-leak B = 2MΩ; − 1·5 V bias	32·4	61·8
Free grid	34·0	58·9
Grid-leak C = 1MΩ; no bias ..	34·5	58·0

observation for nearly three years showed no sign of this condition, that the valve circuit was more likely to be the seat of the trouble. This conclusion may appear somewhat weak when it has been shown that the fork, which is presumably free from these variations, is much more susceptible to circuital variations. The inference is that the transfer of the frequency of operation from 1 to 50 kilocycles per sec. is responsible for the greater frequency instability. Also, occasional variations of greater magnitude than those shown had been noted. These were accompanied by an increase of the anode current and were associated with extreme values of atmospheric humidity. The deduction made from this was that the stray capacity-resistances such as the

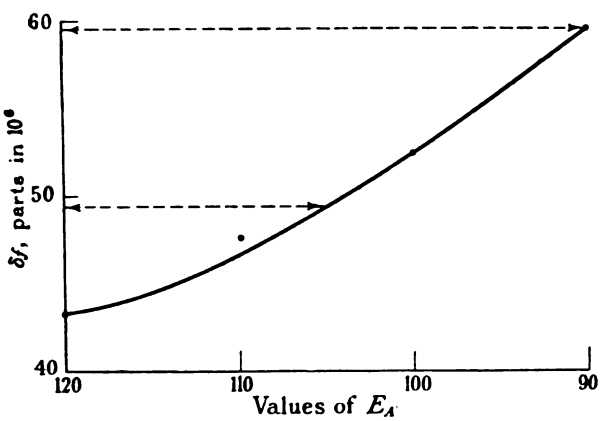

FIG. 17.

valve holder and cap, the two-pin connector of the crystal mount, and the grid-leak mount were varying with atmospheric conditions. These were in shunt with, and affected, the valve input-impedance and, in turn, the resultant frequency.

This deduction was confirmed during the development of the laboratory standard to be described later. In this instance the circuit was as described. A second circuit had the crystal and grid-leak in a desiccating chamber at normal air-pressure. Steps had been taken to reduce to a minimum the amount of dielectric in the remainder of the grid circuit. The circuits in other respects were identical and were exposed to normal atmospheric conditions. The evacuated crystal was, however, much the better oscillator. On one occasion

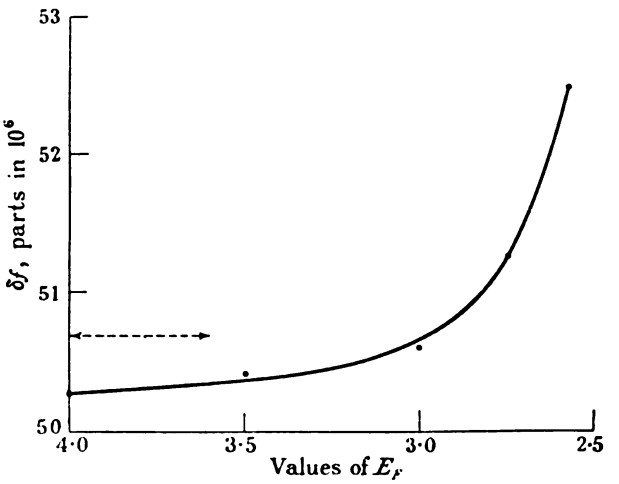

FIG. 18.

during a spell of extreme humidity the evacuated crystal refused to oscillate, while the inferior one in the desiccator oscillated normally. After drying the mount of the evacuated crystal oscillations were obtained with a very high value of anode current and a large frequency change. As conditions improved, the input current

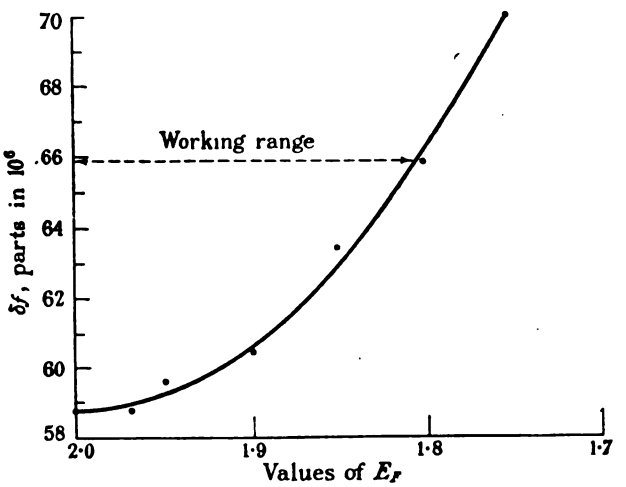

FIG. 19.

and the frequency gradually returned to normal. The other circuit remained constant throughout. This experience was checked by reproducing the conditions artificially. The grid-leak and valve holder, etc., were breathed upon so as to produce a film of moisture and the resultant changes in frequency and anode current were noted. Two tests are recorded in Table 3. The frequency-changes of 30 and 36 parts in 10^6, due to this

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cause, are more than sufficient to account for those previously noted. As the deposited moisture was insufficient to stop oscillations and, indeed, to cause the anode current to reach values which had been observed, it can be stated that this treatment is not drastic and does not exceed the conditions produced by normal atmospheric variations. It must also be pointed out that this phenomenon is quite distinct from that produced by humidity on the crystal itself. It is purely a circuital reaction. It is proposed to investigate the matter quantitatively under controlled humidity conditions, paying particular attention to the grid-current variations. Meanwhile, it was clear that definite precautions would have to be taken in the design, lay-out and mounting of the grid circuit of the valve. Alternatively it would be essential to control the degree of humidity of the whole circuit.

The first condition was partly met by the employment of a new design of L.S.5-type valve, the L.S.5.D. This has the grid connection made to a terminal on the glass envelope, thus eliminating the valve holder and cap from the grid circuit. Also, advantage was taken of some features in the design of the laboratory standard which will now be described.

The Laboratory Standard.

The limit of accuracy to which a crystal, mounted as described, can be adjusted has been found in practice to be of the order of 1 part in 10^4. Also, when once the envelope has been evacuated no further adjustment is possible. The small adjustment of frequency available in the controllable variables is of the order of 1 part in 10^5 and power considerations limit the full use of this. This is satisfactory enough for the purpose for which the instrument was designed—that of a portable standard. It is, however, desirable to eliminate the correction factor for departure from its nominal value when the standard is capable of being adjusted to an accuracy better than 1 in 10^6. To meet this condition, the following controls are available for final adjustment of the frequency:—(*a*) air pressure, (*b*) electrode air-gap, (*c*) anode condenser, (*d*) temperature, (*e*) anode and filament potentials, (*f*) humidity. It was decided to provide for the use of all these.

The design includes a bell jar and desiccator with Fleuss pump, vacuum gauges and an ether-type thermostat. The crystal is mounted directly on the fused-quartz insulating washer of the bell jar with the grid-leak connected to the same terminal inside. The lead from this terminal goes direct to the grid connection on the glass envelope of the L.S.5.D valve. The only dielectric materials in the grid circuit are the valve envelope, the fused-quartz insulating washer, the glass mounting of the grid-leak and the fused-quartz distance pieces of the crystal mount. Of these, the outside of the insulating quartz washer and the valve envelope are the only dielectrics which are subject to normal atmospheric conditions.

The first run on this equipment is shown in Fig. 13 (upper curve). The curve clearly illustrates the effect of humidity upon the frequency of the crystal itself. No oscillations were obtained until 15 minutes after closing the bell jar, and this was not evacuated until the frequency was approaching a steady value. The observed change due to the action of the desiccator is 60 parts in 10^6; to this must be added the change which took place during the period when there were no oscillations. From previous experience on resonators it can be stated that the total change is usually greater than 100 parts in 10^6. It is desired to emphasize this point as the changes due to this variable are, apart from the effects of frictional resistance, greater than those from any other single source. Most current literature makes no reference to this, and it is not understood how the accuracies quoted have been obtained. The change due to the reduction of air pressure to 0·25 in. of mercury is 10 parts in 10^6.

Table 3.

Conditions	Capacity	I_g	I_a	Time for 200 beats	Variation of frequency from 5×10^4 cycles per sec.	Difference
	μμF	mA	mA	secs.	parts in 10^6	parts in 10^6
Normal values for damp weather	145	2·75	7·0	68·4	58·6	0
After breathing on grid circuit	145	3·25	8·5	45	88·8	+30·2
10 minutes later	145	2·5	7·0	66	60·4	+ 1·8
Normal values for damp weather	200	3·5	6·0	65	61·6	0
After breathing on grid circuit	200	2·5	8·2	41	97·6	+36·0
After fanning grid circuit for 2 minutes ..	200	3·5	6·0	64	62·4	+ 0·8

The continuation of this curve gives the readings for the subsequent 60 hours. The large drift of the first 10 hours is due to the further absorption of residual water-vapour by the desiccator. The early runs were taken primarily to investigate the short-period variables, and the thermostatic control was not employed. The remainder of the readings have therefore to be corrected for temperature. The dotted curve shows the corrected readings, and it will be seen that the departure from the mean value is of the order of 2 parts in 10^6. With thermostatic control this figure would be improved, as the room temperatures were used for correction and the crystal lags considerably on these.

The lower curve gives similar readings for the portable-type standard. Also, Fig. 14 gives a continuous-count 20-minute run in the two cases, with observations at

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intervals of 1 minute. The upper curve is interesting as it is apparent that it coincides with a point of maximum departure from the mean, the early portion of the curve just indicating the change from a fall to a rise. The lower curve shows the laboratory standard run and Fig. 12 shows a further run of the same type plotted to a very open scale to permit an appreciation of the observational errors. The data are supplied in Table 4,

TABLE 4.

Time	Reading of chronometer for 120 beats	Time for 120 beats	Variation of frequency from 5×10^4 cycles per sec.	Variation of frequency from 5×10^4 cycles per sec.
1	2	3	4	5
minutes	secs.	secs.	cycles per sec.	parts in 10^6
1	60·6	60·6	1·98	39·6
2	61·6	61·0	1·967	39·34
3	62·6	61·0	1·967	39·34
4	63·5	60·9	1·97	39·4
5	64·4	60·9	1·97	39·4
6	65·2	60·8	1·974	39·48
7	66·2	61·0	1·967	39·34
8	67·2	61·0	1·967	39·34
9	68·0	60·8	1·974	39·48
10	68·8	60·8	1·974	39·48
11	69·8	61·0	1·967	39·34
12	70·5	60·7	1·977	39·54
13	71·4	60·9	1·97	39·4
14	72·0	60·6	1·98	39·6
15	72·8	60·8	1·974	39·48
16	73·8	61·0	1·967	39·34
17	74·4	60·6	1·98	39·6
18	75·4	61·0	1·967	39·34
19	76·0	60·6	1·98	39·6
20	76·8	60·8	1·974	39·48
21	77·4	60·6	1·98	39·6
22	78·4	61·0	1·967	39·34
23	79·0	60·6	1·98	39·6
24	79·8	60·8	1·974	39·48
25	80·4	60·6	1·98	39·6
26	81·0	60·6	1·98	39·6
27	81·6	60·6	1·98	39·6
28	82·4	60·8	1·974	39·48
29	83·0	60·6	1·98	39·6
30	83·6	60·6	1·98	39·6

where, under col. 2, the readings of the chronometer seconds-hand for 120 beats are found, each series being taken consecutively. Col. 3 gives the time in seconds for each of these, col. 4 gives the difference in frequency from 5×10^4 cycles per second, and col. 5 shows the same difference expressed as the number of parts in 10^6.

It is estimated that the maximum probable error in reading the chronometer is 0·4 sec. From the table it is seen that this will introduce an error of 0·2 in 10^6. Fig. 12 shows that the maximum departure from the mean value is 0·13 in 10^6. These deviations are, therefore, probably chiefly observational errors. Over the 30 minutes shown, and dismissing the improbable contingency that they are varying at the same rate and in the same direction, the fork and crystal are constant to a value of the order of 1 part in 10^7.

From these curves it may safely be assumed that the frequency instability characteristic of the portable type is absent in the laboratory type. A number of observations on the latter, covering some 6 weeks, gave variations of the order of 3 parts in 10^6. This is the degree of uncertainty of the reference standard; thus it appeared that the accuracy aimed at (of the order of 1 in 10^6) was probably attained. Also the data were available for the precautions to be taken to obtain a stability of the same order for the portable type. The work of developing the subsidiary apparatus was therefore undertaken.

THE CRYSTAL-CONTROLLED HIGH-FREQUENCY MULTIVIBRATOR.

The features of the well-tried Dye fork-driven multivibrator * indicated the general lines of development. In the first place, the transfer of the operating frequency from 1 to 50 kilocycles per sec. permits the production of the higher frequencies in one step. A multivibrator for 50 kilocycles per sec. was designed and built, the

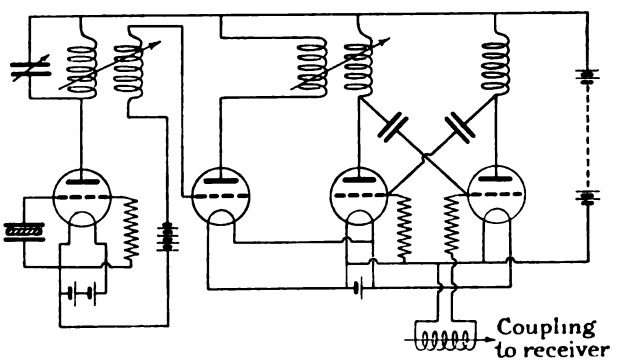

FIG. 20.

only differences from the Dye type being the different value of anode impedances necessary to give 50 kilocycles per sec. and the use of a commercial-type filament rheostat. This is possible because there is ample power in the drive to exercise a wide range of control. There is no need to tune this circuit by an accurate adjustment of the filament potential. The circuit used is as in Fig. 20. For accuracies of 1 in 10 000 the coupling valve may be omitted, but for higher accuracy it should be included. The range of this circuit has not been fully explored at the high-frequency end, but no difficulty has been experienced in working down to the 120th harmonic, i.e. 6 000 kilocycles per sec. A well-constructed open-scale calibrated oscillator is used for harmonic identification. This controlled high-frequency multivibrator was put into service and it proved to be quite satisfactory, but the need for a low-frequency controlled source for interpolation and other purposes was felt.

A further development of this high-frequency multivibrator, however, is needed. Transmission engineers are faced with the fact that current practice of control

* "A Self-contained Standard Harmonic Wave-meter," *Philosophical Transactions of the Royal Society*, 1924, vol. 224, p. 259.

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by transverse oscillators and stable master-drive circuits is not meeting the present needs in the matter of precision and frequency stability. Under ideal circumstances it is possible to produce short-wave transverse oscillators of a frequency within 1 part in 10 000 of the nominal value, provided that temperature, circuital conditions and the crystal mount are all strictly specified. To reproduce the frequency within these limits in a power circuit has proved to be an exceedingly difficult matter. It appears that it will be necessary to employ something in the nature of a standard for the frequency control. This has actually been put into practice at the long-wave G.P.O. station at Rugby. It is becoming an urgent necessity to develop the drive of the multivibrator so that it can be utilized in the higher-frequency bands, of the order of 3×10^5, with its present accuracy when used at 5×10^4 cycles per sec. This will permit the harmonics at extremely high frequencies to be used as " drives " of high precision and stability.

The Quartz-Controlled Low-Frequency Multivibrator.

The idea was conceived of controlling a source of relaxation oscillations of the type produced by the Abraham-Bloch multivibrator circuit by impulsing a harmonic with a forced oscillation. This was tried by injecting an e.m.f. from the crystal-controlled circuit (frequency 50 kilocycles per sec.) into a multivibrator of frequency 1 kilocycle per sec., i.e. impulsing the 50th harmonic. This experiment was undertaken somewhat casually and, to the author's surprise, evidence of perfect control was obvious from the first. No difficulty was experienced in combining this with the high-frequency multivibrator with the drive exercising simultaneous controls. The circuit for this is given in Fig. 21.

This circuit has all the essential features of the standard multivibrator, i.e. a range of 1 to 6 000 kilocycles per second in steps of 1 kilocycle per second, with the addition of the characteristic partial harmonic effects. A minor property is, however, not present. The power to use the 19th, 20th and 21st harmonics of the low-frequency side to control the high frequencies is not available. This property, however, reappears in a more useful form in the centre-drive development. Apart from the 1 kilocycle per sec. of the fundamental of the low-frequency side a number of others on either side of this value are obtainable by adjustment of the coupling,

Table 5.

Harmonic order	Output frequency
	kilocycles per sec.
44	1·135
45	1·111
46	1·086
47	1·063
48	1·041
49	1·02
50	1·0
51	0·981
52	0·962
53	0·943
54	0·926
55	0·909
56	0·983

or of the variable condenser. These are the fundamentals obtained by impulsing successive harmonics. Thirteen of these are available in steps as shown in Table 5. These are

$$\frac{f}{n-6} \cdots \frac{f}{n-1} \cdots \frac{f}{n}, \frac{f}{n+1} \cdots \frac{f}{n+6}$$

where f is the control frequency and n the normal impulsed harmonic frequency, in this case 5×10^4 and 50 cycles per sec. respectively.

An interesting proposal to take advantage of this property has been made by Mr. Williams of the Air Ministry Research Establishment at South Farnborough. He suggests the use of a quartz drive at 50 000 cycles per sec. controlling two multivibrators at frequencies of f/n and $f/(n \pm 1)$, where $n = 100$. The outputs of these multivibrators will appear as 500 and 495 or 505 cycles per sec., respectively. These are combined and the resultant beats of the order of 5 per sec. are counted

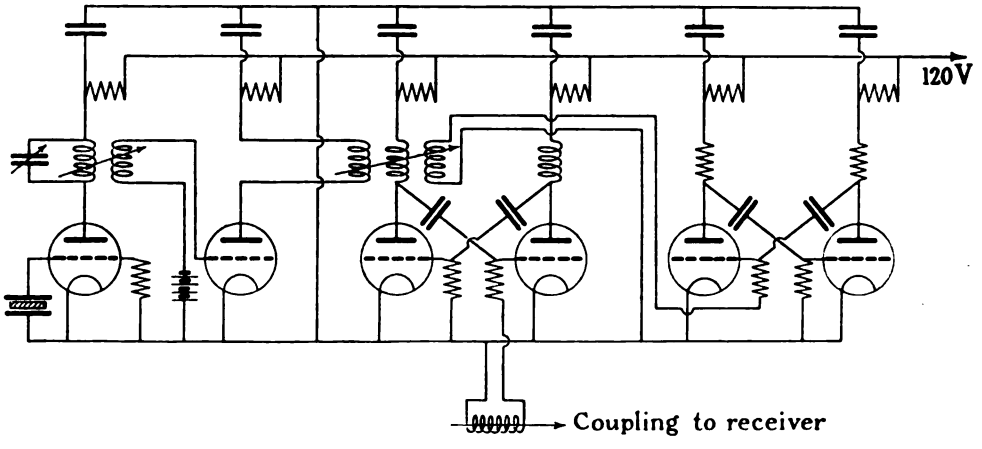

Fig. 21.

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electrically over a period of time by taking photographic records at definite time-intervals. A simple and direct reference to the time standard is thus effected without the intervention of the usual phonic motor or similar device.

This application of the principle of frequency division was made independently and had been in use some weeks when a paper describing a similar application arrived in this country.* This paper describes the control of a multivibrator of frequency 1 kilocycle per sec. by a crystal drive (frequency 50 kilocycles per sec.). The authors state that, although it is possible to exercise control in one step, they actually use two stages. In the present case this has not been found necessary. Four complete sets are in operation, one of which has been in continuous service for over a year, and in all cases complete control has been obtained. The above-mentioned paper gives full evidence of the accuracy of control provided.

The development of this system of controlled precision low-frequency oscillations offers considerable possibilities. No reason is seen why oscillations of a frequency approaching 1 cycle per sec. may not be produced. A direct reference to the ultimate time standard would thus be possible. Indeed, venturing into the debatable realm of prophecy, this may be the germ of the time standard of the future.

Another variant of this method of control is under investigation. It has been found possible to exercise effective control by injecting into a multivibrator forced oscillations having a frequency of $1/n$ times the approximate period of the multivibrator. Thus a 1 kilocycle-per-sec. fork was used to control a 10 kilocycle-per-sec. multivibrator. It is hoped to utilize this principle to refer back to the time standard by impulsing a neon-tube circuit of, say, 10 cycles per sec. from a standard clock and using this in turn to control a 1 kilocycle-per-sec. multivibrator.

It is possible that in the case quoted above the 10th harmonic of the fork is the controlling factor.

* L. M. Hull and J. K. Clapp: "A Convenient Method for Referring Secondary Frequency Standards to a Standard Time Interval," *Proceedings of the Institute of Radio Engineers*, 1929, vol. 17, p. 252.

Trend of Future Developments.

While the equipment thus described more than meets the requirements originally laid down, i.e. the calibration and maintenance of a sub-standard wavemeter to 1 part in 10^4, it is fully realized that this is not the time to call a halt. The question of the behaviour of the quartz standard over relatively long periods of time and, in particular, the question of ageing will require years of observation with equipment of the highest precision. This will necessitate the development of improved methods of reference to the time standards. Also the question of the adequacy of the time standard as the final arbiter of frequency may at any time be questioned.

The experience gained as to the behaviour of the valve and its tendency to develop frequency instability at medium frequencies calls for further investigation, and this is actively proceeding. In this direction it is proposed to lower the frequency to 25 kilocycles per sec.; this represents the limit to which the natural quartz crystal can be obtained and worked. It is to be hoped that the production of large synthetic quartz crystals will not prove such a difficult proposition as the somewhat similar case of the artificial diamond. It is, however, felt that the true solution and urgent need is the development of the resonator reference-standard and its associated equipment to accuracies of the order 1 in 10^6 or, if possible, 1 in 10^7.

The author's thanks are due to the Air Ministry for permission to publish this paper. He also desires to express indebtedness to the inspiration of Dr. Dye's classic work already referred to. Part (1) of the paper is obviously a direct practical development of this work Acknowledgments are due to Messrs. H. W. Sullivan for the loan of the standard valve-maintained fork. Valuable advice has been received from the author's colleagues; in particular from Mr. B. Williams, B.Sc., of the Air Ministry research staff at South Farnborough, and from Mr. R. H. Billingsley, without whose help in checking the experimental work and in the preparation of the paper and drawings a private research of this nature could not have been undertaken in the limited time available.

DISCUSSION BEFORE THE WIRELESS SECTION ON THE PAPERS BY MR. LUCAS (See Page 151) AND MR. J. E. P. VIGOUREUX (See Page 41), 5th MARCH, 1930.

Air Commodore A. D. Warrington-Morris: With regard to the paper by Mr. Lucas, there is no doubt that, with the great increase in wireless communication and the consequent necessity for as many wave channels as possible, the two most important problems which concern wireless engineers at the present time are those of the maintenance of a constant frequency and the improvement of frequency standards. The author has been responsible for inspecting wavemeters for the Royal Air Force, and as a result of those duties he has realized the desirability of a high standard of accuracy. The work described in the paper has been done without the use of laboratory facilities. Table 3, for instance, shows that humidity was varied by the simple expedient of breathing on the grid circuit. Some quantitative measurements of humidity would probably be useful in a paper of this sort. So far as aircraft signalling is concerned, the degree of accuracy referred to by the author is not yet so essential as to warrant the provision in an aeroplane of the additional weight, space and complicated apparatus which it necessitates. For point-to-point services, however, especially on the wave-bands settled for fixed stations by the Washington Convention, the greatest possible accuracy is required. The author appears to view with disfavour the use of a quartz control employing a transverse mode, on account of the high temperature-coefficient. I should like to know whether he has any further objections, and whether he can give some information as to the direct use of the longitudinal mode on short waves without frequency multiplication.

Dr. D. W. Dye: Mr. Lucas has produced a frequency standard of a type superior to any other so far designed

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in this country for frequencies of the order concerned. The problem of transferring them to lower frequencies has been successfully solved in the system devised by Mr. Williams, whereby two neighbouring harmonics are selected and controlled by a multivibrator. That is, in my opinion, a very ingenious arrangement; it is the first of its kind, so far as I know, to be employed in stepping down. It would be somewhat difficult, I think, to keep these systems in step with the crystal over the long periods of time which would be essential in checking absolute frequency measurements against some other

a nodal point of the longitudinal mode of vibration, but not in quite the same way as that adopted by Mr. Lucas; mine was balanced as on a knife-edge on the narrow edge of the crystal. Fig. 5 led me to think that there was something unusual in the values given, but the author's explanation of the errors in scaling explains the apparently remarkable results. Concerning the use of crevasses, this method is most valuable for determining the damping and elucidating the equivalent networks for oscillators or resonators, but it is not the most suitable for determining or setting a source to the precise

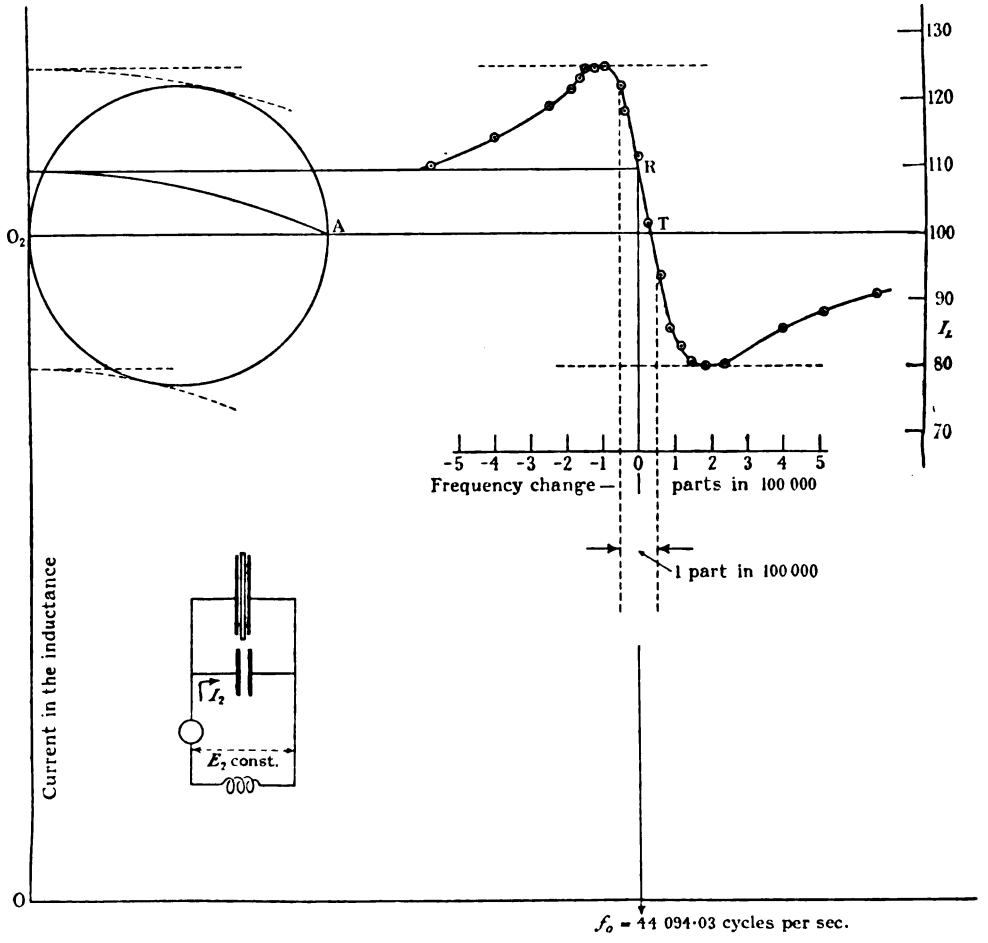

FIG. A.

time standard, but the system is evidently worthy of further investigation and development. With regard to the mounting of the crystal, in my own work on the subject I soon found that humidity had a considerable effect upon the oscillations, particularly in making the damping large when there was much moisture in the atmosphere. I did not find it difficult to get satisfactory behaviour, without using special precautions. It was found that a bell-jar and desiccating agent in the enclosure was sufficient to keep the specimens in a dry condition while being investigated, and to ensure that the results could be reproduced from time to time. With regard to mounting, I a'so have supported the crystal at

frequency of resonance. If the resonator is connected to a small condenser (of, say, 300 $\mu\mu$F) and a practically constant e.m.f. is applied to the combination by the use of a coil of small inductance connected to the condenser, the current in the inductance as a function of frequency is as shown in Fig. A. As the frequency is increased, the current rises to a maximum, turns sharply and steeply downwards to a minimum and then slowly approaches its original value. The frequency of resonance lies on the steeply falling, nearly straight portion, but is not at the point of intersection of this curve with the datum current for frequencies near, but outside, the region of resonance. The graphical construction corre-

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sponding to this case is very simple and is shown on the left of the curve. A circle is drawn of diameter equal to $(\theta_1 - \theta_2)$, where θ_1 and θ_2 are the maximum and minimum currents respectively. This circle has its centre at the height of the current value just outside resonance. The vertical tangent OO_2 is drawn intersecting the abscissa scale at 0. Any point on the curve of current is transferred to the circle by drawing an arc, of radius equal to the current at the point in question, from the centre O. The intersection of this arc on the circle gives the corresponding point on it. The resonance point on the circle is at A and the converse process of transferring this to the current curve is obvious as shown, giving the point R on the curve. The above construction assumes that the air condenser used has no losses. For the mere purpose of setting a frequency, it is not necessary to carry out the graphical construction. If θ_1 and θ_2 are the maximum and minimum currents, the current at which resonance occurs is given by $\theta_r^2 = \theta_1^2 + \theta_2^2 - \theta_1\theta_2$. The resonant frequency may then be found directly by adjusting it until this current is obtained. It is convenient to have a square-law scale on the galvanometer for the purpose. The two vertical dotted lines show a belt of frequency of width equal to 10 parts in a million. A change of deflection of about 30 cm on a galvanometer scale is given for this change. A setting to about 1 part in 10 million can be made if such extreme accuracy is desired. With regard to the frequency changes given in Mr. Lucas's paper, it would be desirable to indicate on the curves the direction in which the change in frequency occurred; this can only be inferred from the text. I think that in Fig. 9 there must be an error either in the scale of frequency changes observed on the broadcast station 2LO, or in the frequency assigned to the base line. The latter is marked as 10 kilocycles per sec., but the range of frequency is from $+ 0{\cdot}5$ to $- 0{\cdot}5$ kilocycle per sec. This means that the variations were of the order of 2 or 3 per cent, which is very unlikely. I wish now to refer to the frequency-changes dealt with in Part (2) of the paper. Mr. Lucas states, when giving the desiderata of the drive circuit, that experience has shown that any attempt to introduce retroaction produces a tendency to frequency instability; and that the circuit should therefore be capable of oscillating without any retroaction apart from that inherent in it. This seems to me to make a distinction without a difference and, although the meaning is clear, I am not certain that the right conclusion has been drawn. It might be advantageous to make use of the good-quality screen-grid valves now available, and then to introduce the retroaction independently, but that is a matter of opinion. I agree with the other two desiderata given, although the statement that the electrostatic coupling of the crystal is much less than that of magneto-striction, fork, and other such electromagnetically driven devices, may require modification. As a general statement it is probably true, but it must not be supposed that finality has been reached in regard to other forms of oscillator and that these could not be designed to have a coupling as small as, or smaller than, that of the quartz oscillator. The question of the small magnitudes of the damping and of the necessary coupling determines the variability of the whole system with regard to the external conditions applied. In Fig. 3 it would be interesting if another curve, starting at the instant of switching on, could be given. My experience with the transverse form of oscillator shows that the energy absorbed in the oscillator itself, although very small, does raise the temperature by 2 degrees or more above that of the electrodes, in the course of half an hour. This causes a drift in frequency which cannot be overcome by any temperature control, because it occurs inside the oscillator itself. As a result of the use of the thin longitudinal bars, the effect might be worse, but the temperature coefficient would be smaller. Thus, if an anode battery or other external factor varies, besides the possibility of a small instantaneous change, there will inevitably be a drift due to the change in the energy in the oscillations of the crystal. It would be interesting to have some figures on this point. Measurements of the variation of frequency due to a small condenser placed across either the grid-anode or grid-filament of the valve would also be welcome. Considerable variations do occur from these causes and, as Mr. Vigoureux's paper shows, frequency variations of 25 to 30 parts in a million are produced by changes in capacity of a few micro-microfarads. These figures apply to higher frequencies, and I assume that the variations would be smaller for a low-frequency oscillator. Mr. Lucas criticizes somewhat adversely the tuning fork which he employed, in which large frequency variations occurred with changes in the circuit conditions, notably in the filament voltage. These variations are much larger than those of the forks which I have used. I have not investigated the forks manufactured by Messrs. H. W. Sullivan, although they were manufactured more or less to one of my own designs. By means of a new magnetic system on the forks I have been able to reduce all these errors to such a small magnitude that a variation of 30 per cent in the anode voltage produces a frequency change of less than 1 part in a million. A change of 10 per cent in filament voltage also produces a frequency change of less than 1 part in a million.

Mr. C. O. Browne: I notice that Mr. Lucas mounts his quartz resonator by means of two pieces of cork situated at the centre of the crystal, presumably at the ends of the electric axis. Can he give any information as to the effect of mounting the crystal in the same way but with the cork at the sides of the crystal? As described in his paper, the resonator is certainly held at a nodal point of mechanical displacement for the fundamental longitudinal mode, but it is also supported at an antinode for vibrations due to the inverse piezo-electric effect. With reference to the curves in Fig. 15, I notice that changes in the anode-condenser setting have been plotted against the resultant changes in frequency. I would suggest that such errors might perhaps be eliminated by substituting a pure resistance for the inductance-capacity circuit connected to the anode of the valve. It is well known that a crystal can be made to oscillate with a resistance only in the anode circuit. If Mr. Lucas has already tried this arrangement, will he state whether the errors due to the resistance are greater than those due to the inductance and condenser combined?

Mr. B. Williams: The need for deriving the fundamental frequency of the crystal from two successive

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harmonics arose from a necessity for checking certain claims in connection with this apparatus. As there was no equipment of the required accuracy available, this method suggested itself to me as a means of measuring the absolute frequency. As Dr. Dye has pointed out, although this is theoretically a simple process it presents difficulties in practice. The problem of the control of low-frequency multivibrators from high-frequency crystals involves a considerable number of variables, some of which have already been tracked down. It necessitates the maintenance of a constant output from the drive, and of steady anode and filament values in the multivibrators. I am chiefly concerned with the practical side of Mr. Vigoureux's paper and regard it as most useful from the point of view of the production of sub-standard quartz oscillators, to accuracies of the order of 1 in 10 000. From this standpoint, the variables associated with the quartz and the valve which can be utilized for the production of minute changes of frequency without going to the ultimate limit of grinding, are of importance. They are especially so in connection with the design of quartz oscillators *in vacuo* from the corresponding quartz resonators. In connection with the difference between the range of capacities for the two methods of using the crystal (i.e. the plate-grid and grid-filament modes), Mr. Vigoureux gives, in equation (73), an expression for this range in terms of the anode-grid capacity, and states that for the other mode of vibration the two ranges of capacity are in direct proportion to the inter-electrode capacities. By noting the values of these changes of capacity, therefore, it should be possible to measure the grid-anode capacity of a screen-grid valve. For instance, if Mr. Vigoureux's constants for the quartz and the valve are used in equation (73), but instead of 2·4 $\mu\mu$F a value of 0·005 $\mu\mu$F is substituted for the grid-anode capacity, the actual change of capacity is found to be 2 $\mu\mu$F. Such a quantity is difficult to measure, but it can be measured. On testing this in practice with a screen-grid valve I obtained values at 50 000 cycles per sec. which were considerably larger than this. I should like to ask Mr. Vigoureux whether the equations would hold for very small values of the inter-electrode capacities. Would the method which I have outlined be suitable for checking the capacities of the screen-grid valve?

Mr. J. E. P. Vigoureux : In Mr. Lucas's paper there is a remark about the accuracy to which the response frequency of a quartz resonator can be determined from the crevasse curve. It may be worth mentioning that there is a certain ratio of minimum to maximum ordinate of the curves for which the accuracy of the frequency determination is greatest. The curves are obtained by measuring either the voltage across the tuning circuit or the current in the inductance of the latter, and if the deflection of the recording instrument is plotted against the frequency a curve similar to that shown in Fig. 10 of the paper results. Dr. Dye has pointed out that if the curvature at its lowest point were a maximum, the location of the response frequency would be as accurate as possible. The minimum can always be altered by changing the ratio of capacity to inductance, or by adding to or reducing the resistance of the tuned circuit. If the deflection of the recording instrument be proportional to the square of the current or voltage measured, best conditions for frequency measurement are obtained when the minimum ordinate is half the maximum ordinate.

Mr. H. J. Lucas : I have been unable to do more than concentrate on those portions of Mr. Vigoureux's paper immediately bearing upon the frequency instabilities which I have been investigating. The conclusion at which I arrived—that these were caused by variations of the valve-input impedance with atmospheric humidity—receives certain confirmation. The frequency changes produced by variations in grid conductance and grid-filament capacity, as shown by Mr. Vigoureux, are in the same sense as those produced by humidity variations. The magnitude of the g_2 variations is, however, too small to account for the observed departures and the C_2 variations only partially account for them. I do not find Mr. Vigoureux's argument in favour of the grid-anode connection very convincing. I submit that the question of frequency stability is of more importance than any consideration of power output. The majority of the curves for this connection show that any given change of circuit constants produces a larger frequency change than in the grid-filament connection. It is of interest to note that the crystal used for this investigation has two well-defined response frequencies and is capable of generating these frequencies simultaneously. This is a common phenomenon of transverse oscillators and is a serious handicap to the production of good crystals for control purposes. A very able paper by F. R. Lack * discusses the point and arrives at the conclusion that it is a characteristic of the " parallel " cut crystal. From the values given by Mr. Vigoureux I deduce that his crystal is a " perpendicular " cut. My own experience also confirms the fact that the latter cut suffers from this disability. Mr. Vigoureux's paper is already proving very useful in solving current problems, and I have no doubt that it will prove invaluable as a standard work of reference for a considerable period to workers in this field.

Mr. W. H. F. Griffiths (*communicated*): I wish to congratulate Mr. Lucas on a paper that conveys no real idea of the vast amount of work which he must have carried out. Only those directly interested in the standardization of electrical quantities can have any idea of the difficulties which are encountered in such work and which can only be overcome by much patience and resource. For this reason one must be careful not to speak lightly of accuracies higher than, say, 1 part in 10 000. As an admirer of Dr. Dye's past work on the development of a valve-maintained tuning fork as a frequency standard, I am naturally at present in favour of the fork-controlled multivibrator system as a standard of radio frequencies, more especially in view of the present improvements which he is effecting both in the fork itself and in the associated equipment for checking and watching the variation of its frequency. The portable crystal standard described in the paper, however, undoubtedly fills effectively the gap immediately below this standard. When viewed as a very precise sub-standard, second only to the fork, its temperature coefficient is entirely negli-

* " Observations on Modes of Vibration of Quartz Crystal Plates," *Proceedings of the Institute of Radio Engineers*, 1929, vol. 17, p. 1123.

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gible while the fault inherent in the portable standard—the inability to adjust the frequency more closely to its nominal value than 1 part in 10 000—is rendered less important, as corrections for accuracies higher than this will only be needed occasionally. The stability appears to be excellent for such a purpose. Very precise sub-standards are absolutely necessary if full use is to be made of standards having accuracies of a few parts in 10^6. Realizing this, I have been working upon the problem in a different way by endeavouring to rely upon the calibration permanence of inductance and capacity in a simple resonant circuit, and the value of sub-standards of this kind would be increased if an occasional check of capacity "drift" were made with the portable crystal reference standard described by the author. By designing special fixed inductances and variable condensers I have found it possible to produce a wide-range wavemeter (it is a comparatively simple matter to obtain accuracy in sub-standard wavemeters of limited range) with a temperature coefficient of the order 1 part in 10^5 per degree C. and with a degree of calibration permanence of the order 1 part in 10^4 over considerable periods of time. The comparatively negligible temperature-coefficient is due to a thermal compensation principle employed in the construction of the inductances, and the high degree of calibration permanence is due to a dimensional compensation principle employed in the construction of the variable condenser which renders it as permanent as a fixed air condenser of first quality. Because of its simplicity and rapidity of operation I feel that this type of continuously variable sub-standard will have a useful field of application, especially if an accurate quartz resonator is used at periods of a few months to give an indication of the ageing of the variable condenser at its maximum capacity setting. At this setting the variable condenser approximates to a fixed air condenser of minimum dielectric gap and maximum plate area, and the greatest percentage capacity change is therefore expected in this position. I should be glad to have the author's views and suggestions upon this application of the reference standard. I should perhaps add that for frequencies lower than 3 000 kilocycles per sec. I have found it possible to increase the utility of this sub-standard by converting it into a generator. This is done by associating its simple resonant circuit with a negative resistance device (i.e. without reaction, in the manner suggested by Dr. N. W. McLachlan) in the shape of a 4-electrode valve in which the inter-electrode capacity is reduced to a value such that a slight relative displacement of the electrodes will not affect the accuracy of the wavemeter. To suit the purpose, a valve has been produced by collaboration between Dr. McLachlan and the General Electric Co., Hammersmith. Used in this manner the wavemeter is practically independent of quite large changes in the potentials of the electrodes of its valve and of the filament current of the latter. Its stability over short periods is such that the frequency has remained constant to 2 or 3 parts in 10^6 for 12 hours in a room of varying temperature, the check being made by short-period comparisons against the 220-kilocycle harmonic of a multivibrator controlled by the fork standard used by Mr. Lucas and in a manner similar to that described on page 157.

Mr. H. J. Lucas (*in reply*): I wish to thank Dr. Dye for his suggestion of a method of adjusting the frequency of an oscillator to that of a resonator. I consider it doubtful whether any normal oscillating circuit will be sufficiently stable to enable full advantage to be taken of the accuracy of adjustment which the method allows, but I hope to use a quartz-controlled oscillator checked at intervals by comparison with a resonator, using the controllable variables for adjustment. I agree that it would have been better to have indicated the direction of the various changes in frequency, but this can be deduced from the text and the fact that both crystals were approximately 2 cycles per sec. below their nominal frequency. The scale errors in Figs. 5 and 9 have been corrected for the *Journal*. It is interesting to note that desideratum (3) is receiving confirmation in recent improvements in the valve-maintained fork. In view of the improvements indicated by Dr. Dye the advent of the new standard tuning fork will be received with great interest.

With regard to the question of the employment of retroaction, it was considered that the feed-back by the valve inter-electrode capacity, besides being sufficient for all normal purposes, should be a stable quantity inasmuch as it could only vary by movement of the valve electrodes, which would naturally be guarded against. A screen-grid valve was not employed, no bombarded valve of this type then being available, and it was anticipated that variations in grid contact potential would be troublesome in a "gettered" valve.

I regret that time does not permit of experiments being conducted on the rise of temperature of a crystal under oscillatory conditions.

I made no criticisms of the tuning fork during my work and I accepted it as my frequency standard without question. No evidence appeared to suggest that the fork had any inaccuracy greater than that anticipated, that is, 2 or 3 parts in a million under operating conditions when the supply voltages were maintained at their correct values.

I fully concur that quantitative measurements of humidity would have proved valuable, but the conditions under which the work was carried out precluded this. The point raised by Air Commodore Warrington-Morris regarding the use of the transverse mode of vibration for control purposes is interesting, but the objections are (1) the indicated reason—high temperature coefficient, (2) the production of spurious frequencies in the neighbourhood of the fundamental due to the coupling of this and other modes of vibration, and (3), as Dr. Dye has shown, the impossibility of obtaining nodal support for a transverse oscillator. It has not been found possible to make satisfactory longitudinal crystals having a frequency above 300 kilocycles per sec.

An instance of the use of a resistive load referred to by Mr. Browne has been given by Hull and Clapp.* I considered that either method of supporting the crystal, i.e. at the ends of or at right angles to the electric axis, introduces a definite amount of damping by operating on the reciprocal vibration, and this would probably account for the decrement of my best evacuated crystal being no better than that of Dr. Dye's crystal in air.

* See reference on page 163.

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It is hoped to employ the information provided by Mr. Vigoureux in future work on resonators.

Mr. J. E. P. Vigoureux (*in reply*): Replying to Mr. Williams's inquiry as to the validity of formulæ (42) and (73) for very small values of the inter-electrode capacities, these formulæ are obtained, as pointed out on page 46, on the assumption that the range of capacity in the plate circuit, over which oscillations can be maintained, namely $C \sim C_0$, is fairly large. If the plate-grid capacity is very small, this range will also be small, and formula (73) is not applicable. It is then no longer true to say that " the total range of capacity C over which oscillations are possible with the filament-to-grid connection is to the corresponding range in the plate-to-grid connection as C_3 is to C_2." But an idea of the range can be obtained from formula (71) by assuming usual values for the various constants of the valve. I find that with $C_3 = 8$ $\mu\mu$F, $m = 100$, $g_0 = 10^{-5}$ mho, and the other variables approximately the same as those given in Table 1 of the paper, the range for the plate-grid connection is just over 4 000 $\mu\mu$F. With the filament-grid connection and values of 0·1, 0·08 and 0·07 $\mu\mu$F respectively for the plate-grid capacity, the ranges would be 41, 24 and 12 $\mu\mu$F. When the plate-grid capacity is 0·06 $\mu\mu$F or less, oscillations cannot be maintained. If those figures represent anything like the actual state of things, it would seem that the oscillations observed by Mr. Williams were due to imperfect elimination of capacity between the plate and the grid leads, outside the valve.

A few experiments were carried out to check the theoretical results obtained above. The quartz disc the constants of which are given in the paper was used again, with an air-gap of 0·012 cm. The amplification factor of the valve was 180, its plate resistance was 240 000 ohms, and the plate and screen-grid voltages were 115 and 84 respectively. A 1·2-megohm grid leak was used. At first the oscillatory circuit was made up of a shielded variable air condenser, a 1-ohm vacuo-junction and an ordinary cylindrical 160-μH coil. When no particular precautions were taken to eliminate capacity between the plate circuit and that electrode of the resonator connected to the grid, oscillations took place fairly strongly. By merely screening the lead connecting the plate to the insulated terminal of the condenser, the amplitude of the oscillations and the range of capacity over which they could be maintained were considerably decreased.

A toroidal coil of inductance 100 μH was next wound and enclosed in a copper box; the leads connecting it to the vacuo-junction and to the insulated terminal of the condenser were also screened, and all the screens were maintained at filament potential, either by direct connection or by large condensers. With this arrangement it was impossible to obtain any oscillations. One end of a screened lead was then connected to the plate terminal. To the other end was soldered a thin brass plate 3 × 7 cm, the distance of which from the resonator electrode joined to the grid could be adjusted at will It was thus possible to add a small capacity between the plate and the grid, although the actual value added could not be estimated in these preliminary experiments. It was then possible to obtain oscillations, provided the added capacity exceeded a certain value. For the limiting value, and for two larger values of this capacity, the ranges were 20, 30 and 47 $\mu\mu$F, and the amplitude of the maximum difference of potential across the toroidal coil was less than 3 volts in the first case; it was 7 in the second and 17 in the third. A 5-$\mu\mu$F shielded condenser between the grid and the plate increased the range to 430 $\mu\mu$F and the amplitude to 35 volts.

These results show that the plate-grid capacity is too small to maintain oscillations, and it seems that the method cannot be employed, without modifications, to measure the plate-grid capacities of screen-grid valves.

On page 46 of the paper, second line above equation (42), F should be replaced by E, and the ordinates of Fig. 6 should be multiplied by 10 000.

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RECENT DEVELOPMENTS IN DIRECTION-FINDING APPARATUS.*

By R. H. BARFIELD, M.Sc.(Eng.), Associate Member.

(*Paper first received 8th December, 1929, and in final form 3rd March, 1930; read before the* WIRELESS SECTION *2nd April, 1930.*)

SUMMARY.

Part (1) describes a four-aerial direction-finding system erected at a site near the Radio Research Station, Slough, and gives details of the tests carried out on this apparatus at wavelengths of 300 to 600 m. The system is based on the Adcock principle, but a modification has been introduced to correct for the asymmetry of the aerial system with respect to its upper and lower portions which is unavoidable if the operating hut is at ground-level.

In the first test the performance of the new direction-finding system was compared with that of a rotating-loop direction-finder over a prolonged period which included conditions of most severe "night effect." In the second test a calibration of the system was made by means of a transmitter attached to a kite which provided a source of downcoming waves of known angle of incidence and polarization.

The results of the tests show the new system to be considerably superior to the closed-loop system for conditions of "night effect." They indicate, however, a residual instrumental error of the system in the presence of downcoming waves. Finally, a further modification of the system by which it is hoped this will be eliminated is described.

Part (2) describes two distinct types of portable short-wave direction-finding apparatus and gives details of the methods by which they were tested and compared under working conditions at wavelengths of 12 to 60 m.

The first instrument is of the totally-screened closed-loop type.

The second instrument is of the rotating Adcock type consisting of a single pair of spaced vertical aerials pivoted about a vertical axis.

The tests described are divided into four sections:—

(*a*) Tests for instrumental error with a local transmitter situated first on the ground and then elevated 40 ft. above the ground with its aerial at various angles to the vertical to give waves of varying polarization.

(*b*) Tests on the effects of the immediate surroundings of the receiver.

(*c*) Direction-finding observations on the "ground ray" of a transmitter up to the maximum distance at which the ray is detectable.

(*d*) Long-range tests at which only downcoming waves were being received.

Test (*a*) showed the Adcock to be much superior to the loop in the presence of downcoming horizontally polarized waves, while (*b*), (*c*) and (*d*) show that for observations on the ground ray both instruments are equally satisfactory and that with due precaution in the choice of a site their performance is much the same as that of longer-wave apparatus. For observations just outside the range of the ground ray both systems are equally useless, while at greater ranges (beyond the "skip distance") the results again demonstrate the superiority of the Adcock to the loop.

It is shown that the results generally confirm the conclusions with respect to the nature of the propagation of the waves arrived at by Eckersley† from somewhat similar experiments.

*Reprinted from *Journal I.E.E.*, 1930, vol. 68, p. 1052.

†T. L. ECKERSLEY: "Short-Wave Wireless Telegraphy," *Journal I.E.E.*, 1927, vol. 65, p. 600; also *Proceedings of the Wireless Section*, 1927, vol. 2, p. 85.

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Part 1. MEDIUM WAVES (300 TO 600 M).

(1) INTRODUCTION.

It is well known that direction-finding apparatus based on the closed-loop principle such as the Bellini-Tosi single- or double-coil systems—in fact all existing systems which are employed in a practical capacity as an aid to marine or aerial navigation—are subject to a very serious inherent limitation or defect. They determine not the direction of travel of the waves but only that of the resultant horizontal magnetic field of the waves. When it is a question of taking measurements on the ground ray of a wave alone these two directions are at right angles, but when, on the other hand, it is a question of receiving waves which have travelled via the Heaviside layer, some of which arrive from the upper atmosphere with horizontally polarized components, there is no necessary relationship between them. In fact these systems are entirely unreliable as direction-finders under such conditions.

The first attempt to produce an instrument not subject to this limitation was made in 1916 by Adcock,‡ who pointed out that a directional receiving system consisting of spaced vertical aerials should theoretically determine the true direction of travel under all circumstances. Adcock actually constructed such an apparatus

‡ F. ADCOCK: "Improvement in Means for Determining the Direction of a Distant Source of Electromagnetic Radiation," British Patent No. 130490 (1919). *See also* R. L. SMITH-ROSE and R. H. BARFIELD: "The Cause and Elimination of Night Errors in Radio Direction-Finding," *Journal I.E.E.*, 1926, vol. 64, p. 831; also *Proceedings of the Wireless Section*, 1926, vol. 1, p. 145.

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and demonstrated the improved nature of the results to be obtained with such a system, though this work was not actually published until after the lapse of several years.

During this time attempts were being made at Slough to produce a direction-finder free from the defect described above, which might be conveniently designated " horizontal effect "—a direction-of-travel-finder as distinct from a direction-of-field-finder.

A system based on Adcock's principle was in due course produced which gave almost complete freedom from " horizontal effect." The system, however, relied for its success on the suspension of the hut containing the operator and apparatus at a height of some 20 ft. above the ground. An account of this work is given in an earlier publication by Dr. Smith-Rose and the author.

The subsequent work described in this paper has been devoted to the development of the Adcock principle as a practical system for medium waves by removing the drawback of an elevated hut, and to the construction of portable apparatus (described in Part 2). A considerable amount of attention has been devoted to testing and calibrating.

(2) Description of Balanced Adcock System.

The system differs from that described in the earlier paper* in three respects:—(1) The hut containing the apparatus is on the ground, (2) balancing condensers are

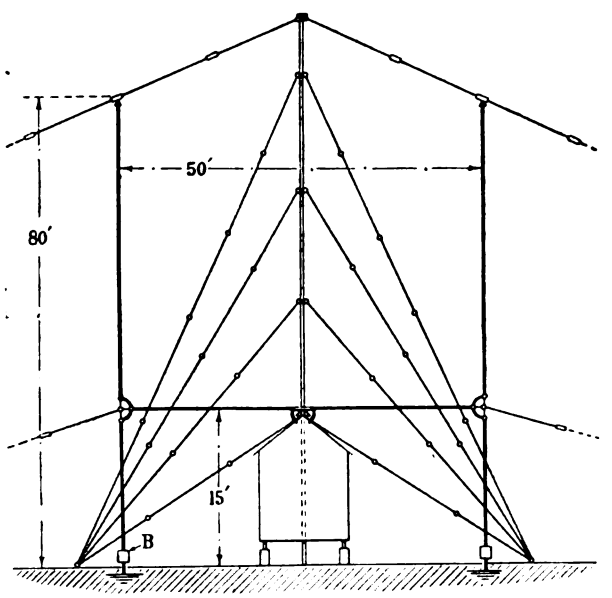

Fig. 1.—Aerial system.
B = box containing balancing condenser.

inserted in the earth leads to compensate for the lowering of the hut, and (3) the aerial system is supported by a single central mast instead of four separate masts (see Fig. 1).

Operating Hut.—The hut is 15 ft. square and is thoroughly screened by means of 1-inch mesh wire netting with which the outside is covered. The 100-ft. mast passes up through the centre of the hut in order to ensure an equal effect on all four aerials.

Aerial System (*see Figs.* 1 *and* 2).—The four aerials approximately 80 ft. high are suspended from four radial stays of stranded steel which are anchored to the ground at points 400 ft. out from the mast. These stays are " broken " by suitable insulators into 20-ft. lengths.

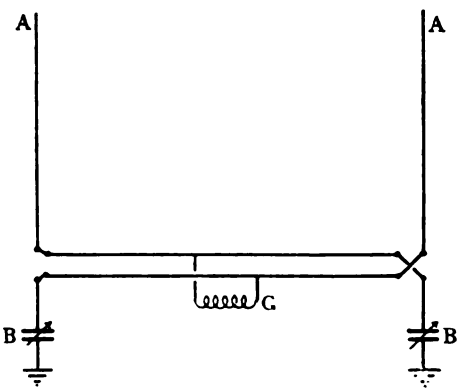

Fig. 2.—Diagram of aerial connections.
G = field coil of goniometer.
B = balancing condensers.
A = aerial.

The system of stays supporting the mast is also broken up by insulators into lengths not exceeding 15 ft. Each aerial is stretched tight to anchor pegs in the ground vertically under its point of support, and is broken at a point 15 ft. above the earth by an insulated length containing the spacing element for the horizontal leads.

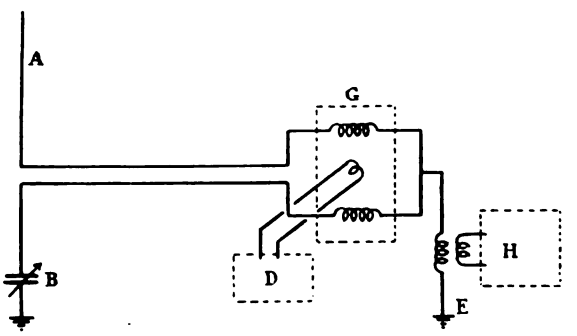

Fig. 3.—Method of obtaining balance.
A = single aerial of Adcock system.
B = balancing condenser.
D = amplifier, etc.
E = earth of centre of system.
G = goniometer.
H = local high-frequency source.

Horizontal Leads.—The twin horizontal lead-in wires are attached to the insulators on the spacing piece in the aerial wire, which keeps them 6 in. apart in a horizontal plane. They are strained to insulators attached to battens on the mast at a point just above the roof ridge of the hut. Connections are made to the aerial by soldered " tags." A separate down-lead is attached to each of the eight horizontal wires and these are

* *Journal I.E.E.*, 1926, vol. 64, p. 831; also *Proceedings of the Wireless Section*, 1926, vol. 1, p. 145.

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passed through the roof of the hut by means of lead-in insulators.

Earth Connections.—Each earth connection consists of four cylindrical holes 4 ft. deep and 2 ft. apart, arranged symmetrically round the foot of the aerial. Each hole contains a piece of metal sheet bent in the form of a cylinder to fit the hole. The holes are then filled in and rammed down. The four wires are "commoned" and soldered to the earth terminal of the balancing condenser.

Balancing Condensers (*see Fig.* 2).—These are variable condensers* (maximum capacity 0·0003 μF) of commercial pattern entirely enclosed in a metal "plymax" box (10 in. × 10 in. × 8 in.) which is fixed to wooden pegs

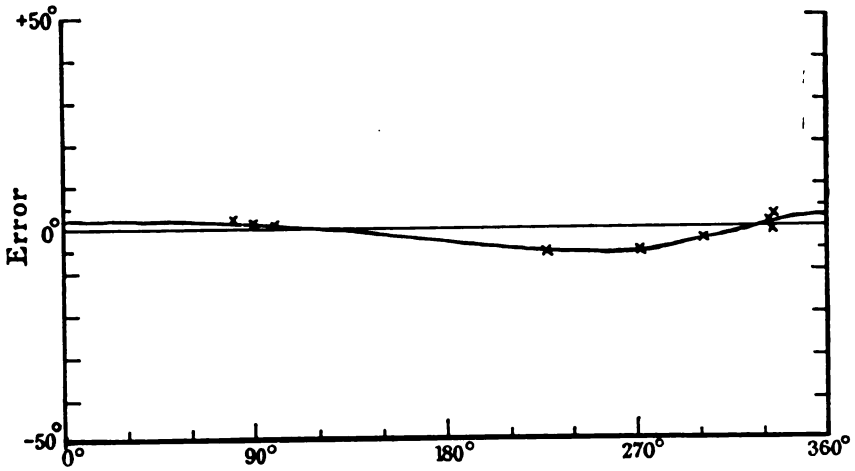

FIG. 4.—Calibration curve for ground waves.

TABLE 1.

Comparison of Standard Deviation for Balanced Adcock and Direction-Finding Loop.

Date	Time (G.M.T.)	Station	D.F. loop		Adcock (Bal.)	
			Number	S.D.*	Number	S.D.*
3. 5.28	19.15–20.15	Bournemouth	28	30·6	81	10·44
10. 7.28	18.40–21.30	Bournemouth	65	29·2	73	1·1
11. 7.28	19.15–21.50	Bournemouth	100	25·9	65	0·5
17. 7.28	19.15–22.00	Daventry	37	0	41	0·12
19. 7.28	19.30–22.00	Daventry	87	1·51	59	0·08
24. 7.28	19.00–21.25	Manchester	87	14·95	118	7·4
25. 7.28	19.45–21.30	Manchester	72	33·1	59	3·3
15.10.28	17.10–18.50	Bournemouth	—	†	88	8·8
14.11.28	17.25–18.25	Daventry	84	0·72	50	0·08
27.11.28	17.15–18.30	Langenberg	120	9·1	85	3·2
29.11.28	16.53–19.00	Langenberg	218	12·5	151	3·0
11. 1.29	18.30–19.05	Hamburg	44	0·97	35	1·25
23. 1.29	17.30–18.30	Hamburg	102	1·7	43	0·7
28. 6.29	20.45–21.15	Glasgow	45	4·6	24	0·93
11. 7.29	20.15–21.30	London	121	0·62	60	0·29
18. 7.29	20.15–21.30	London	117	0·63	71	0·3
9. 8.29	19.25–20.00	Cardiff	45	25·6	47	3·95
15. 8.29	19.10–20.00	Cardiff	40	19·46	48	2·23
22. 8.29	19.40–20.20	Dublin	60	3·84	58	1·98

* S.D. = standard deviation. † Night error too severe to take standard deviation (see Fig. 5c).

Note:—The standard deviation is obtained by taking the root-mean-square value of the variation of individual observations from their arithmetical mean.

* These condensers may be replaced by fixed condensers as soon as the correct value has been found: they do not have to be readjusted on changing wavelength unless the natural wavelength of the aerial is approached.

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driven into the ground. A short-circuiting switch is mounted on the outside of the box. When the aerial system is correctly balanced these condensers have a value of approximately 60 $\mu\mu$F.

Interior Connections.—The interconnections between the horizontal leads necessary to complete the arrangement shown in Fig. 2 are made inside the hut at roof-level. Each lead, however, is broken at the point of entry into the hut by means of a switch controllable from the ground by means of a rod. The four down-leads to the goniometer are brought down the sides of the mast.

Goniometer and Receiver Apparatus.—These do not differ in any important way from the installation of a standard Bellini-Tosi direction-finding system.

Balancing the System.—As explained later, the object of the balancing condensers is to make the impedance of the lower half of the system equal to that of the upper half with respect to e.m.f.'s generated in the horizontal members of the aerial system by the horizontal electric component of the received waves. This is achieved on each aerial separately by the device shown in Fig. 3. Each of the four aerials in turn is connected to its earth through both field coils of the goniometer in the manner shown, and the point of junction of the field coils is connected to an earth at the centre of the aerial system; actually the screening wire of the hut is good enough. A local carefully-screened high-frequency source is set going to inject an e.m.f. into this centre earth-lead, thus sending a current which branches off into each field coil. One branch passes along one horizontal lead into the aerial, and the other passes along the other horizontal lead to earth through the balancing condenser. The relative intensity of these

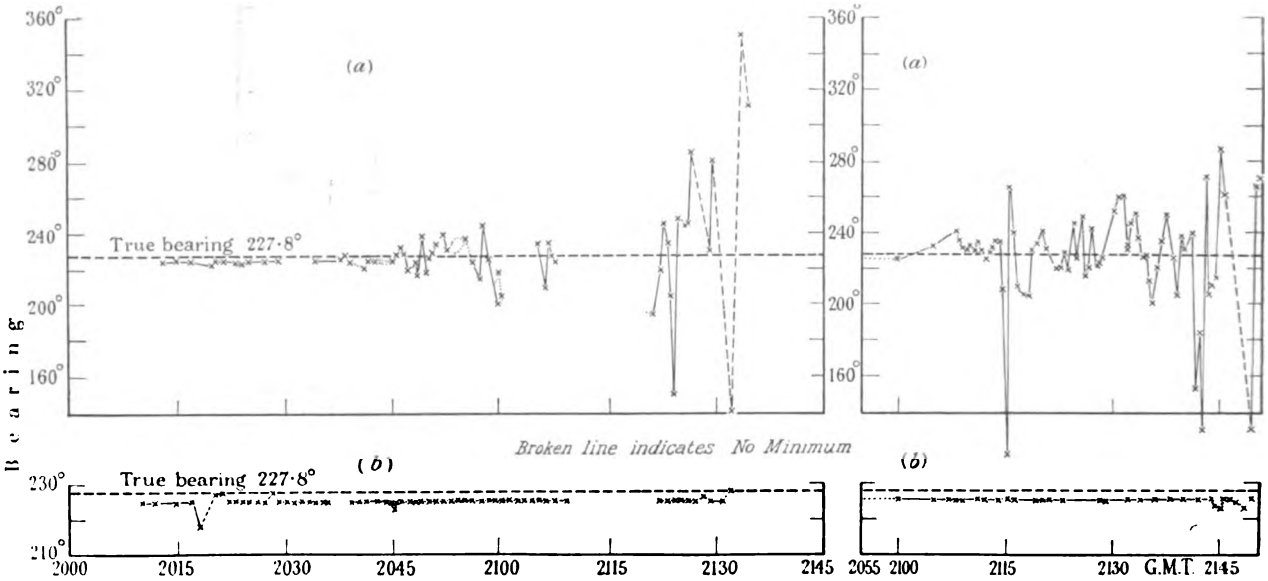

FIG. 5A.—Comparison of simultaneous bearings on (*a*) direction-finding coil system, and (*b*) Adcock system. Transmitter: Bournemouth (329 m). Left-hand 10.7.28; right-hand 11.7.28.

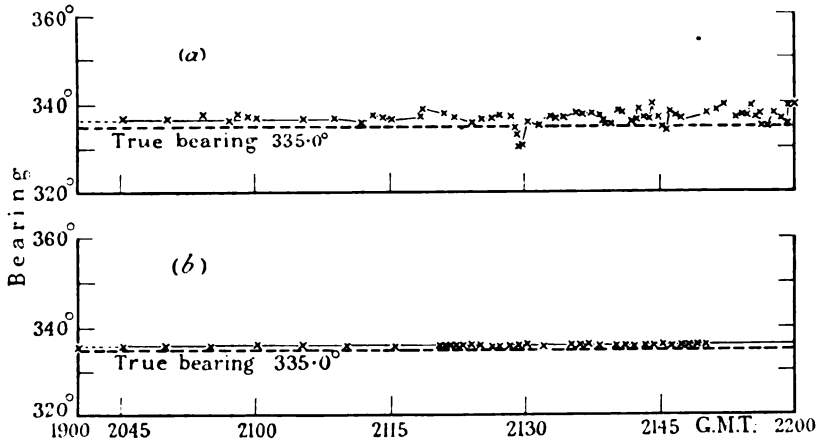

FIG. 5B.—Comparison of simultaneous bearings taken on (*a*) direction-finding coil system, and (*b*) Adcock system. Transmitter: Daventry (491 m), 18.7.28.

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two currents can now be determined from the position of the search coil necessary to obtain a " minimum " from this local signal, and by adjusting the balancing condenser until this minimum occurs with the search coil at 45° to each field coil they are made of equal intensity. The system is then balanced.

(3) TESTS.

The apparatus was subjected to the following tests:—

(*a*) Calibration for ground waves and range tests.
(*b*) Tests during " night effect."
  (i) Comparison with closed-loop direction-finder.
  (ii) Effect of balancing condensers.
(*c*) Calibration by means of downcoming waves from kite transmitter.

A description of these tests, with their results, is given in this section.

(*a*) *Calibration for Ground Waves and Range Tests.*—This test was carried out in the usual manner by taking bearings of a number of distant transmitting stations in as many different directions as possible. Broadcasting stations in the United Kingdom were for the most part observed and the observations were made in the daytime when ground waves only were being received. The calibration curve is shown in Fig. 4 and compares favourably with those obtained from other forms of direction-finding systems under these conditions. No difficulty was experienced in obtaining sharp minima. The range of the installation with standard forms of amplifier appeared to be not noticeably different from the Bellini-Tosi or other system.

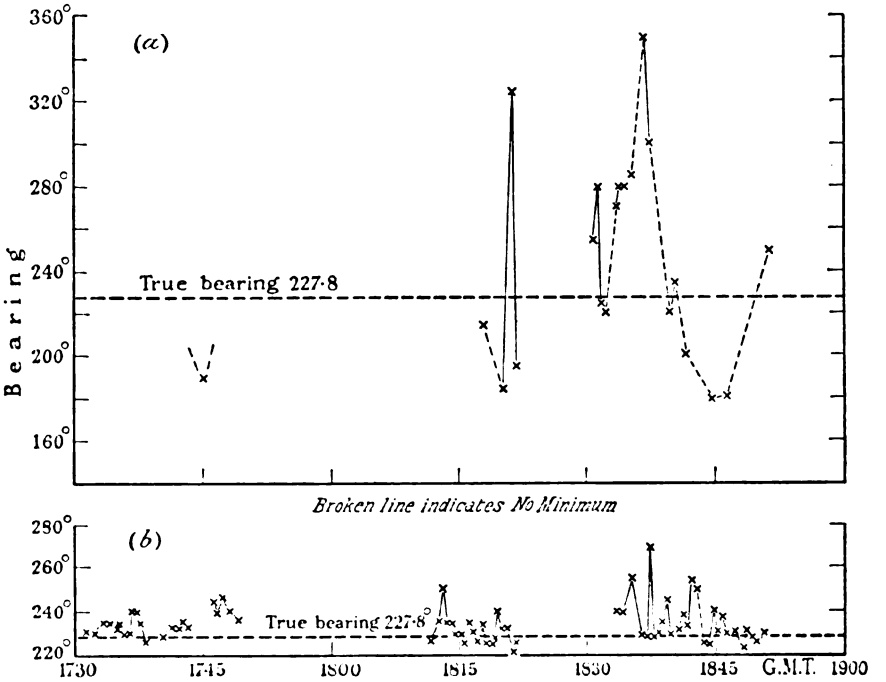

FIG. 5C.—Comparison of simultaneous bearings taken on (*a*) direction-finding coil system, and (*b*) Adcock system. Transmitter: Bournemouth (329 m), 15.10.28.

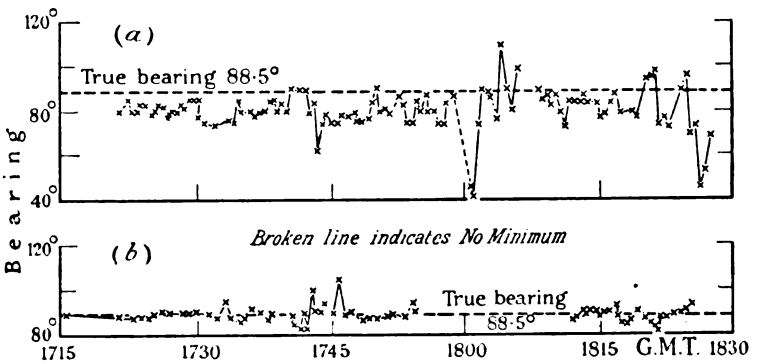

FIG. 5D.—Comparison of simultaneous bearings taken on (*a*) direction-finding coil system, and (*b*) Adcock system. Transmitter: Langenberg (468·7 m), 27.11.28.

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(*b*) *Tests during "Night Effect."*

(i) A series of observations of bearings of distant transmitting stations extending over a period of about 18 months have been made in which the performance of the Adcock system, as described in this paper, was compared with that of a rotating-coil direction-finder under conditions of "night effect." Most of the observations were taken over a period of one or two hours just after sunset.

Table 1 gives a complete summary of all the results obtained during these tests. It gives the standard deviation for each period of observation and for each type of direction-finder. The curves in Figs. 5A, 5B, 5C, 5D, and 5E show the observed bearings plotted with respect to time of observation in a few selected cases chosen to include both the best and worst from the point of view of the Adcock direction-finder.

Both the tables and the curves show in an unmistakable manner the superiority of the Adcock system over the loop system, and little further comment is necessary. It should be pointed out, however, that in cases where the night effect is very severe, this method of presenting the results tends to underestimate the magnitude of the disturbance on the loop, since, in these cases, no reading whatever can be obtained on the loop over large portions of the period of observation. With the Adcock, on the other hand, such lapses are so extremely rare that the curves and the table give a just account of its performance. It should further be borne in mind, when considering the Adcock results by themselves as an indication of its suitability for a practical direction-finding system, that the particular circumstances in these tests were deliberately chosen to be those of the worst possible "night effect" obtainable. Thus, for example, on longer waves or over sea, nothing so marked as the state of affairs indicated by Fig. 5c is ever experienced.

(ii) Separate tests were made to prove the efficacy of the method of "balancing" introduced in this system. These tests consisted in taking alternative sets of observations on the Adcock only, of about ¼-hour duration each, with and without the balancing condensers in the earth leads (see Fig. 2). The results are shown in Fig. 6 and in Table 2, and clearly indicate the improvement obtained by the use of the condensers.

Returning to a consideration of Fig. 5, it will be noticed that most of the curves show a residual error of a few degrees obtained on the Adcock system, while in one case (Fig. 5c) during the occurrence of exceptionally bad "night effect" on the loop direction-finder

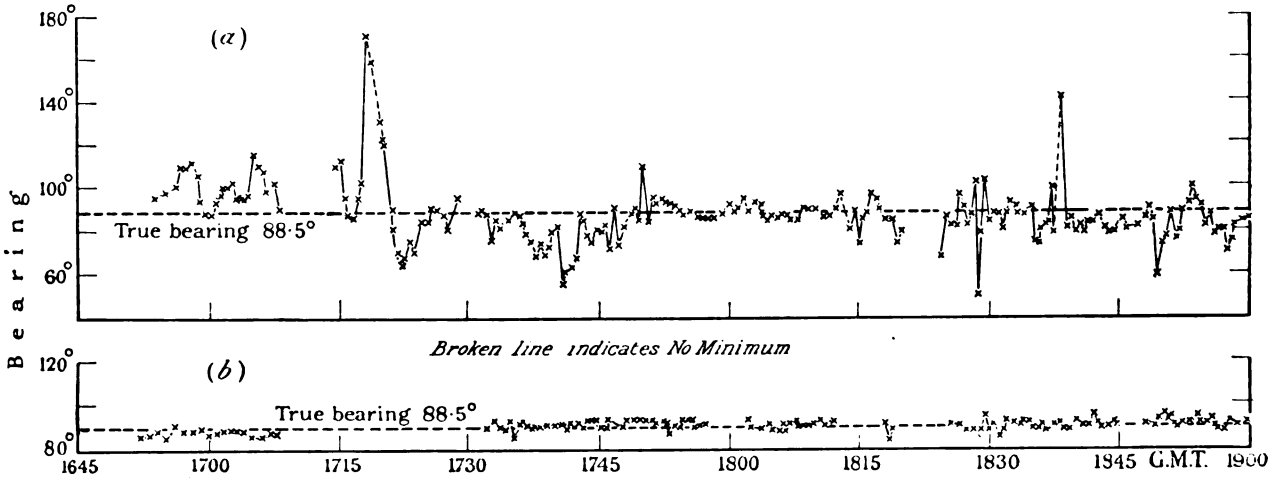

FIG. 5E.—Comparison of simultaneous bearings taken on (*a*) direction-finding coil system, and (*b*) Adcock system, 29.11.28.

TABLE 2.

Comparison of Standard Deviation for Balanced and Unbalanced Adcock.

Date	Station	Unbalanced		Balanced	
		Number	S.D.*	Number	S.D.*
10.10.28	Bournemouth	5	9·52	27	5·77
25.10.28	Bournemouth	16	10·29	37	2·07
1.11.28	Bournemouth	20	28·1	49	8·1
25.7.28	Manchester	51	14·9	59	3·3

* S.D. = standard deviation.

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the maximum error of the Adcock was as high as 40°. The question naturally arises whether these errors are instrumental and curable or whether they are due to actual lateral deviation and therefore incurable. Some advance towards answering this question is given by the tests with artificially produced downcoming waves, which will now be described.

(*c*) *Calibration by means of Downcoming Waves from Kite Transmitter.*—In order to arrive at the cause of

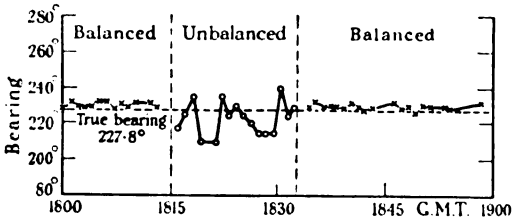

FIG. 6.—Effect of balancing condensers.

the residual errors experienced in the Adcock system during " night effect," it was necessary to calibrate the system with a source of artificially produced downcoming waves with a known and suitable degree of horizontal polarization. These waves were produced by means of a transmitter raised several hundred feet above the receiver by means of a kite. The transmitter consisted of the simplest form of valve oscillator directly coupled to a light flexible aerial wire 100 ft. long, twisted in with the string of the kite. The arrangement is shown diagrammatically in Fig. 7.

In order to reduce weight the high-tension and the valve filament-current were supplied from small dry cells of the grid-bias or flash-lamp type. The capacity of these cells was enough to give at least 6 hours' running, and this was enough for the purpose of one test. The whole transmitter weighed about 2½ lb. The kite was borrowed from the Meteorological Office and is the type used for carrying up scientific instruments. With the complete arrangement it was possible to work with the transmitter 200 or 300 yards away from the Adcock aerial, and with waves arriving at an angle of incidence varying from 80° to 30°. The inclination to the vertical of the transmitting aerial, which corresponded to the angle of the kite string, varied from about 30° to 60°, and this varied the angle of polarization of the waves in a corresponding manner.

A complete set of measurements consisted in a simultaneous determination of:—(1) The true bearing (ϕ) of the transmitter, (2) the apparent bearing (A) as observed on the Adcock system, (3) the angle of incidence (θ) of

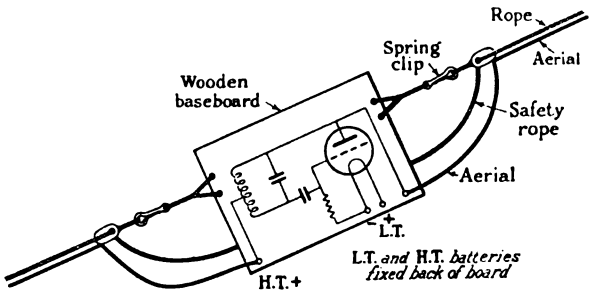

FIG. 7.—Kite transmitter.

the downcoming waves, and (4) their angle of polarization (α). Of these, ϕ was measured with sufficient accuracy by means of a sighting arm pivoted about a vertical axis and provided with a pointer and scale, and θ and α by sighting through a transparent protractor fitted with a plumb bob.

In Fig. 8A are shown the results of these tests, the error (ϕ — A) being plotted against the angle of polarization of the transmitter for various angles of incidence. On each curve is also plotted the corresponding error for the closed-loop direction-finder, as calculated theoretically and verified experimentally. A careful study of these curves, which should be compared with those in Fig. 17, gives a clear impression of the extent to which each system is affected by abnormally polarized waves, and shows again the superiority of the Adcock system. To take one example, Fig. 8A(*c*) shows that a wave arriving with an angle of incidence of 75° (15° to the horizontal) and polarized with its electric force at an angle of 60° to the vertical, will give rise to an error of 24° on the loop direction-finder but only 7° on the Adcock.

On the other hand these kite experiments show that residual experimental errors do exist, so that, considering the two tests together, it would appear that the residual

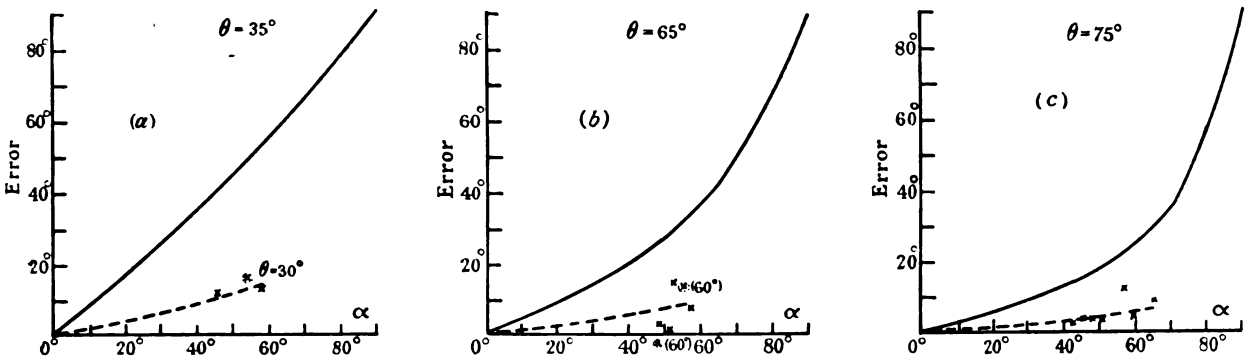

FIG. 8A.—Adcock direction-finder kite tests.

θ = angle of incidence of downcoming wave.
α = inclination of transmitter aerial to vertical.
— curve for closed-loop direction-finder.
- - - curve for Adcock direction-finder.

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errors experienced at night are likely to be due to a predominatingly large extent to instrumental error and are therefore curable. This conclusion is of some

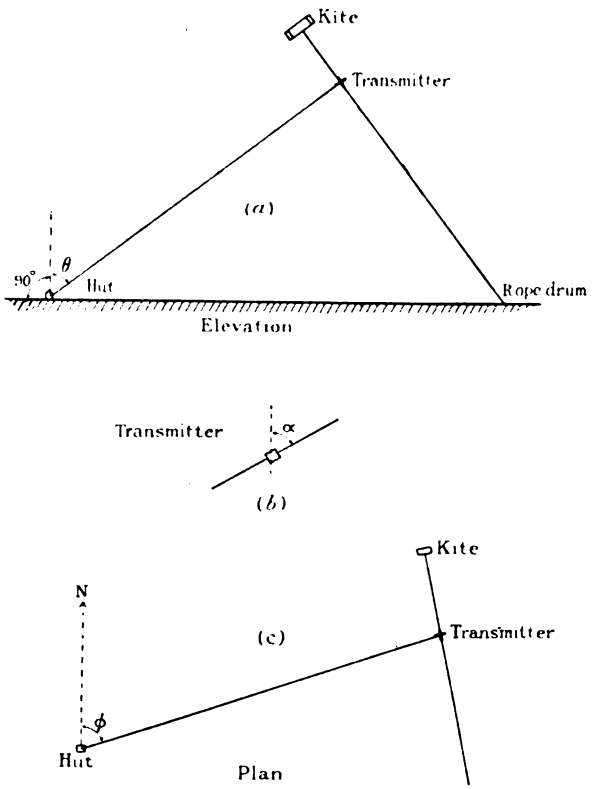

FIG. 8B.

importance as it points to the desirability of a careful theoretical consideration of the system with a view to its further successful development.

induce no current through the field coil "a b." This implies in the first place that the impedance Z_1 of the circuit "r m a b h t" shall be equal to that (Z_2) of the circuit "c n b a g f." If the aerials are untuned with natural wavelength well below that of the working wavelength, this condition is achieved by making the capacity of the balancing condensers equal to that of the aerial circuits. It is interesting to note that the values of the condensers found by the practical method of balancing described above confirm this.

Let I_Y = current through field coil due to Y, due to abnormally polarized wave,
Y = abnormally polarized electric component,
α = horizontal magnetic component at right angles to Y, and
Z_3 = impedance of circuit "c n b a g f."

Then we have for the current through "a b"

$$I_Y = \frac{Y_s}{Z_1} - \frac{Y_s}{Z_2} + \frac{\omega a s h_1}{c Z_3}$$

when Z_1, Z_2 are the impedances defined above and h_1 is the height of the horizontal portion above the ground. By balancing up the condensers we make $Z_1 = Z_2$ but we are still left with the component produced by the magnetic field of the downcoming wave in the loop "c n b a g f," and, since the ratio α/Y varies with the angle of incidence of the wave and the magnetic components are not necessarily in phase with the electric components, it appears impossible to balance this out.

It would at first appear that this component could be eliminated by making h very small, i.e. by having the horizontal leads very close to the ground. Unfortunately, however, in doing this, the capacity to earth of the horizontal limbs becomes no longer negligible, with the result that the balancing condensers are seriously "short-circuited" by distributed capacity to earth along the wire. There is probably, however, a

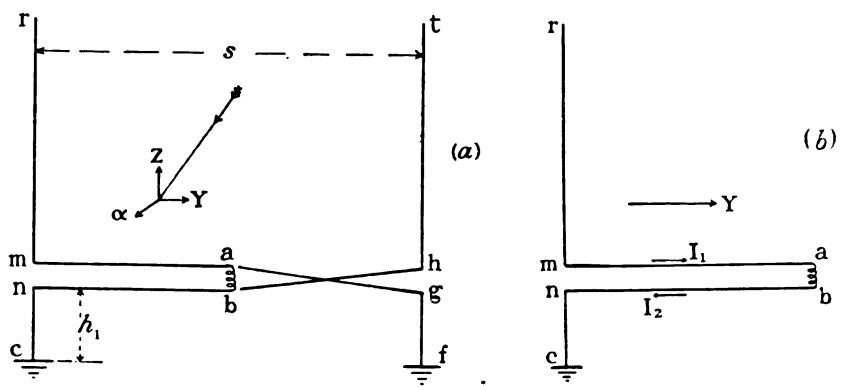

FIG. 9.

THEORY OF BALANCED ADCOCK SYSTEM.

The circuit shown in Fig. 9(a) represents a balanced Adcock "element" under the influence of a downcoming wave with horizontal polarization. The condition for correct balancing of the system is that the horizontal electric component of the received field shall

most favourable height which could be found by a series of careful experiments or possibly by calculation.

We have considered the balancing of the circuit "r m a b h t" against the circuit "c n b a g f." It remains to consider the behaviour of the circuits "r m a b n c" and "t h b a g f," both representing possible paths for currents generated by the Y com-

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ponent. Fig. 9(*b*) shows one of these circuits drawn separately.

It would at first appear that, so far as this circuit is concerned, since the *Y* component induces equal and opposite e.m.f.'s in the two limbs "m a" and "n b," no current due to this field component would flow through the field coil "a b" under any conditions. This would only be the case if the circuit under consideration were a closed circuit.

In an open circuit, owing to the dominating importance of "displacement currents" which originate, so to speak, at all points on the aerial and result in an unequal distribution of current over the aerial, it is clear that equal e.m.f.'s introduced into the circuit at different points will give rise to unequal currents at any specified point in the aerial. Hence the *Y* component, although inducing equal and opposite e.m.f.'s in "m a" and "n b" respectively, actually will in general give rise to a residual resultant current in the circuit which passes through the field coil and is thus a possible source of errors. It will be noted, however, that equal e.m.f.'s in "m a" and "n b" will induce equal currents if the current distribution curve is symmetrical about "a b," i.e. if, when the aerial is normally excited, the current maximum occurs at the centre of the field coil. Fortunately, this can be achieved by inserting a balancing condenser in the earth lead just as before, and the value of the balancing capacities required in the two cases appears to be approximately the same.

It is extremely difficult to calculate the extent of the error produced under any given circumstances, but the following cases give some idea of their importance.

Case 1. *Unbalanced System.*

(*a*) *Effect of Magnetic Component* (*a*).

Let I_Z = current through field coil produced by normally polarized field (*Z*),
I_a = current through field coil produced by abnormally polarized magnetic component (*a*),
X_a = impedance of aerials, and
X_b = impedance of loop.

Then
$$\left.\begin{aligned} I_Z &= \frac{\pi s}{\lambda}\cdot\frac{Zh}{X_A} \\ I_a &= \frac{\omega a s h_1}{c X_b} \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad (1)$$

If δ be the error in degrees we have (neglecting phase differences)

$$\tan\delta = \frac{I_a}{I_Z} = \frac{a}{Z}\cdot\frac{X_A}{X_B}\cdot\frac{h_1}{h}\cdot\frac{\omega\lambda}{c\pi} \quad . \quad . \quad . \quad (2)$$

Take $h_1 = 10$ ft.; $s = 50$ ft.; $h = 60$ ft.;

$X_A = 1000\omega$; $R_A = 30$ ohms; $X_B = 400$ ohms (approx.)

Practical experiments have shown that a/Z frequently exceeds the value $a/Z = 0{\cdot}5$. It is found from (2), using this value, that errors exceeding 10° may result from a system of this type.

(*b*) *Effect of Electric Component.*—The first effect of the electric component referred to is not readily calculable, but the second effect is more manageable and this may be termed the "current distribution effect."

Neglecting the mutual capacity of the two horizontal leads in Fig. 9(*a*) it is permissible to assume sinusoidal current distribution along the aerial according to a law

$$I = K\sin\frac{2\pi x}{\lambda}$$

where x is the distance of the point under consideration from the top of the aerial and K is a constant depending on the height of the aerial and on the amount of series inductance.

It is now necessary to make the assumption that the residual current induced by the *Y* component is equal to the difference between the *mean currents* in horizontal limbs produced by a "lumped" e.m.f. of the same value acting alone in the middle of each limb, or symbolically that

$$I_Y = K\left(\sin\frac{2\pi x_1}{\lambda} - \sin\frac{2\pi x_2}{\lambda}\right)$$

where x_1 and x_2 are the respective distances of the centre points of the horizontal limbs from the aerial tip, and the constant K is given by

$$K\sin\frac{2\pi h'}{\lambda} = \frac{Ys}{2X_A} \quad . \quad . \quad . \quad . \quad (3)$$

where h' = total length of aerial.

As before,
$$I_Z = \frac{\pi s}{\lambda}\cdot\frac{Zh}{X_A}$$

so that

$$\tan\delta = \frac{I_Y}{I_Z} = \frac{K\left(\sin\frac{2\pi x_1}{\lambda} - \sin\frac{2\pi x_2}{\lambda}\right)}{(\pi s/\lambda)\,(Zh/X_A)} \quad . \quad (4)$$

Since $\lambda >> h$ and $x < h$ we may write sines equal to angles, so that, substituting for K in (4), we obtain

$$\begin{aligned} \tan\delta &= \frac{(Ys/2X_A)\,(\lambda/2\pi h)\,.\,\left[2\pi/\lambda(x_1 - x_2)\right]}{(\pi s/\lambda)\,(Zh/X_A)} \\ &= \frac{Y}{Z}\,\frac{x_1 - x_2}{h'}\cdot\frac{\lambda}{2\pi} \\ &= \frac{Y}{Z}\cdot\frac{\lambda}{2\pi h}\cdot\frac{x_1 - x_2}{h'} \quad . \quad . \quad . \quad . \quad . \quad . \quad (5) \end{aligned}$$

It is possible to evaluate this as follows:—

$\lambda = 300$ m; $h = 20$ m; $h' = 40$ m;
$x_1 = 16$ m; $x_2 = 7$ m; $x_1 - x_2 = 9$ m.

Former experiments show that Y/Z may frequently exceed the value $Y/Z = 0{\cdot}5$.

Hence
$$\tan\delta > 0{\cdot}5\cdot\frac{300}{2\pi 20}\cdot\frac{9}{40}$$

which shows that this effect under the particular circumstances may give rise to errors exceeding 14°.

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Case 2. Balanced System.

Effect of Electric Component (Y).—If the balancing has been properly carried out the effects of the Y component should now be negligible either on the circuits "r m a b n c" and its pair or on the circuits "r m a b h t" and "c n b a g f."

Effect of Magnetic Component.—We are, however, still left with the magnetic component acting on the loop "c n b a g f." The error from this source is given as before by

$$\tan \delta = \frac{\alpha}{Z} \cdot \frac{X_A}{X_B} \cdot \frac{h_1}{h}$$

The value of X_A for the aerial system in question has already been given as about 1 000 ohms. The value of X_B is obtained from the value (C_B) of the balancing condensers, and the inductance (L) of the loop "c n b a g f" and that of the field coil. The inductance of the loop is of the order 8×10^{-5} henry; the capacity of the condensers is of the order $1{\cdot}0 \times 10^{-10}$ farad, making the total impedance approximately 2 500 ohms.

Hence $$\frac{X_A}{X_B} = \frac{1}{2{\cdot}5} \cdot \frac{h_1}{h} = \frac{1}{6}$$

Again taking the maximum value of α/Z to be $0{\cdot}5$, we have

$$\tan \delta = 0{\cdot}029$$

and the possible error is less than 3°.

This theoretical comparison of the balanced Adcock with the unbalanced Adcock shows the latter to be liable to errors of much greater magnitude than the former, and is, in general, in agreement with the experiments.

(4) Conclusions and Future Work.

Considering the work as a whole, it is seen that the type of direction-finder described, which may be called a balanced Adcock direction-finder, constitutes a simple practical form of direction-finder which, compared with present existing practical types, has greatly improved characteristics with regard to downcoming waves. In other words it constitutes a system which, although not more complicated than other systems or more difficult to operate, is far less subject to night effect or to "aeroplane effect." The actual extent of the improvement is well represented by the kite tests shown in Fig. 8 and the night-effect tests of Fig. 5, while the degree to which this improvement is due to the addition of the balancing condensers is shown by Fig. 6.

Both theory and experiment, however, show that the system is not entirely free from what may be termed "avoidable error," that is, error caused by the polarization of the waves as distinct from the unavoidable error which would be caused by their actual deviation from the straight path.

Future development work clearly lies in the direction of removing the residual error which is still inherent in the system, i.e. in devising a system for which the curves of Fig. 8 show zero error for all values of θ and α.

Now an extension of the theory outlined above to other forms which the Adcock principle may take shows that any unbalanced Adcock system in which the vertical aerials are connected to the goniometer by

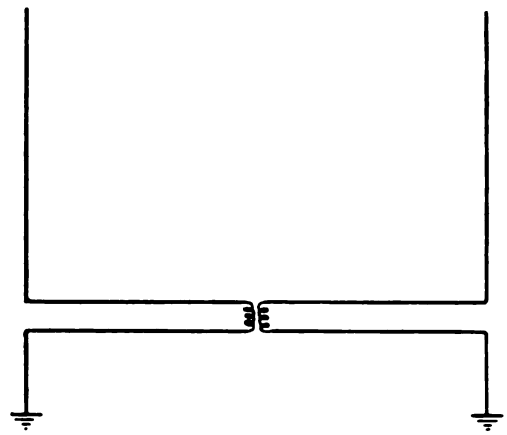

Fig. 10.

continuous horizontal leads is liable to errors resulting from the action of horizontally polarized waves.

Four ways may be distinguished in which this action may take place:—

(*a*) The action of the horizontal electric field on a long horizontal pair of conductors which are in circuits of different impedance, but both in series with the field coil of the goniometer.

(*b*) The action of the horizontal electric field on a long horizontal pair of conductors forming part of the same circuit containing a field coil, but in which the current distribution is not uniform.

(*c*) The action of the horizontal electric field on a

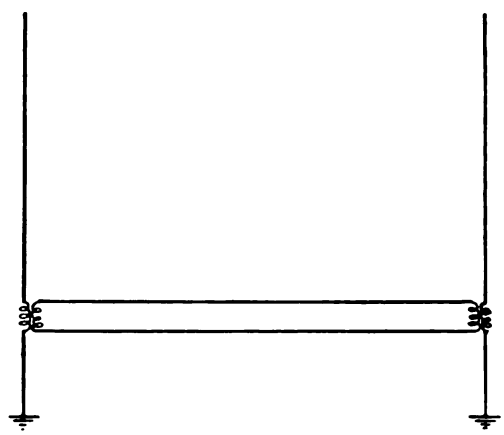

Fig. 11.

long pair of horizontal conductors giving rise to currents in the vertical aerials by virtue of the *electrostatic* coupling between the horizontal conductors and the vertical aerial.

(*d*) The action of the horizontal magnetic field on the closed loop formed by a single conductor, the ground and the lower parts of the vertical aerials when this closed loop contains a field coil.

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These effects are not all present in any given system. At least three types of circuit exist which come under the above classification of an unbalanced Adcock system, and each behaves differently in this respect:—

(1) *Direct-coupled Type.*—This is as shown in Fig. 3 of Adcock's patent specification,* in which the aerials

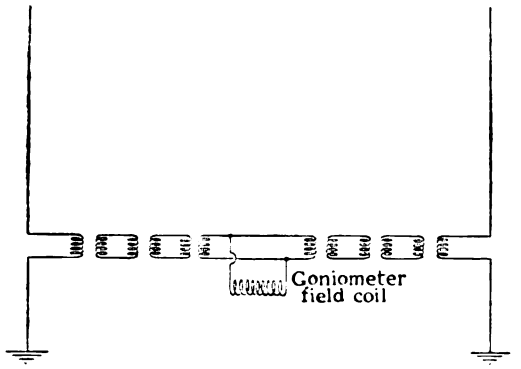

FIG. 12.—"Chain"-coupled Adcock direction-finder.

are directly connected to each other and to the field coil of the goniometer. This is subject to errors of classes (*a*), (*b*) and (*d*).

(2) *Loose-coupled Type* (*Fig.* 10).—In this the aerial circuits are magnetically coupled to the field-coil circuits, the coupling coils being at the centre of the system. This is subject to errors of class (*b*) only.

(3) *Loose-coupled Type* (*Fig.* 11).—In this the arrangement is the same as (2) except that the coupling coils are situated near the aerials. This is subject to errors of class (*c*).

Thus each type is subject to error of some form. In every case it is the horizontal leads which give rise to the error, and although balancing by condensers

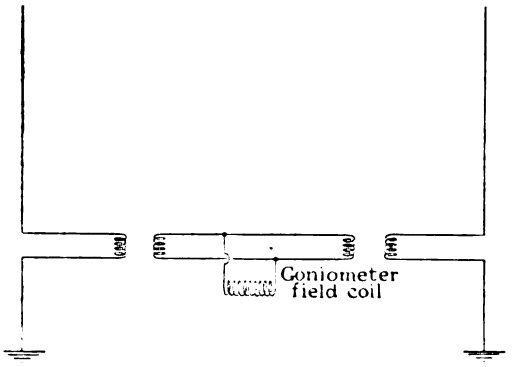

FIG. 13.—Simplest arrangement of chain coupling.

results in great improvement it has been demonstrated in at least one case to leave residual imperfections. It therefore appears that a perfect system would demand the abolition of these leads altogether. This may be virtually achieved by substituting for these leads a series of magnetically coupled circuits, each circuit occupying a comparatively short length of space between the aerial pairs. Fig. 12 illustrates the general principle. The actual number of "links" in the chain here shown has no special significance; theoretically the greater the number the more perfect the system becomes. Fig. 13 shows a particular arrangement when the new principle is introduced in its most practical form. Here the number of links is reduced to two for the sake of avoiding technical difficulties.

Development of the Adcock system therefore involves in the first instance the testing of this principle and this will be attempted if occasion arises.

* *Loc. cit.* (same as Fig. 2, without balancing condensers).

Part 2. SHORT WAVES (12 TO 60 M).

(1) DESCRIPTION OF CLOSED-LOOP DIRECTION-FINDER.

The loop consists of a single conductor forming a square with a side of 3 ft. This conductor is stretched tight at the centre of an enclosing brass tube of 1 in. square section. The 3 ft. square frame formed by the brass tube is electrically continuous except at the centre of the top side, where there is a gap $\frac{1}{8}$ in. wide, the frame being kept rigid at this point by a small ebonite junction-piece fitting tightly into the ends of the tube.

The mechanical construction is shown in Fig. 14. The frame is welded to a hollow brass vertical axis A

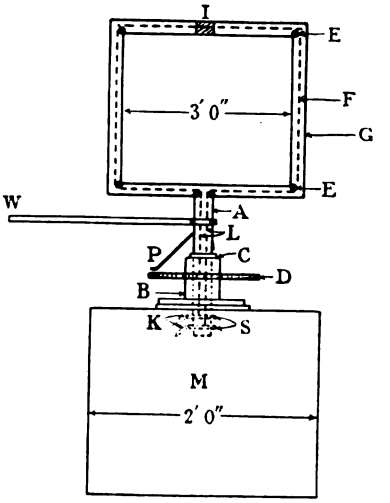

FIG. 14.—Rotating-loop direction-finder for short waves.

A = brass vertical axis $1\frac{1}{2}$ in. diam.
B = brass bearing 3 in. internal diam.
C = brass collar.
D = aluminium scale.
E = ebonite spacing-pieces.
F = direction-finding loop.
G = outer brass pipe.
I = insulated gap.
K = brushes.
L = leads.
M = metal-lined box with amplifiers, tuner, etc.
P = pointer.
S = slip-rings.
W = wooden handle.

$1\frac{1}{2}$ in. diameter which rotates in the fixed brass bearing B and carries the pointer P. The leads L from the single-turn loop are brought down the interior of this axis and terminated on the slip-rings S. Before this is done the wire is strained tight so that it remains rigid in all parts of the tube. At the corners of the brass frame the wire passes over ebonite spacing-pieces E, which allow the wire to slip over them when it is tightened.

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The bearing B consists of a brass cylinder with bottom flange of 3 in. inside diameter and 8 in. long. The inside of the cylinder forms the bearing for the vertical axis, while the outside of the cylinder is machined to take the circular aluminium scale-plate D (8 in. dia.) which is marked off in degrees. This can be rotated relatively to B by undoing a locking screw on occasions when it is required to set the scale to read true bearings. The bearing B is bolted to the centre of the top of a double metal-lined box (plymax) 2 ft. cube and a hole is cut in the top of the box sufficiently large to allow the slip-rings on the vertical axis to pass through. The metal-shielded loop is erected by simply placing the axis in the bearing B so that it rests on the collar C. The lower part of the axis with the slip-rings then projects about 3 in. down into the interior of the box and makes contact with the brushes K, which are "sprung" aside while the loop is lowered into position. The lid of the box is at the side and has a 1-in. overlap. In order to ensure good contact, however, two clamps are mounted on each of the four sides of the box and are screwed up when the lid is in position. The frame is rotated by means of a light wooden handle W clamped to the vertical axis. The slip-ring brushes are connected to the terminals of the coupling coil which forms the input of the amplifier.

The amplifier is mounted on a wooden shelf about 1 ft. from the top of the box, the battery being placed beneath. It consists of a 6-valve super-heterodyne receiver of commercial pattern, the valve filament being supplied from a 2-volt cell. A single-stage magnifier unit is also employed. The single tuning condenser is controlled externally by a suitable extension rod. The receiving loop remains untuned, being simply loose-coupled to the inductance of the rectifying and oscillating valve of the super-heterodyne. Heterodyne reception of continuous-wave stations is effected by allowing the intermediate-frequency stage to oscillate and controlling it to maximum sensitivity by means of a potentiometer. The receiver as a whole is exceptionally simple to operate, there being only one tuning control. The circuit diagram is given in Fig. 15.

The advantages and suitability of this instrument for short-wave working lie mainly in the form of screening employed, and are:—

(1) Complete absence of antenna effect, resulting in very sharp and accurate bearings.
(2) Robustness of loop, since it is enclosed in a metal pipe.
(3) Loop may be approached and, on the higher wavelengths, the screen even touched without appreciably altering bearing or weakening signals.

(2) Description of Rotating Adcock Direction-Finder.

The system consists of two spaced vertical aerials fixed at the ends of a rotating horizontal arm, the aerials being connected in opposition as shown in Fig. 16(*a*). Each aerial is made up of two 5-ft. lengths of $\frac{3}{4}$-in. dia. aluminium tube. The horizontal arm is in two 5-ft. lengths of $1\frac{1}{2}$-in. dia. brass tube, the two lengths being joined at the centre by a brass sleeve or union piece [Fig. 16(*c*) and (*d*)]. In a later model a single length of tube is used for the arm. The twin horizontal leads connecting the two aerials together are stretched tight in the brass tube at a distance of $\frac{3}{4}$ in. apart and at the centre are brought out to two terminals for connecting to the primary of the transformer.

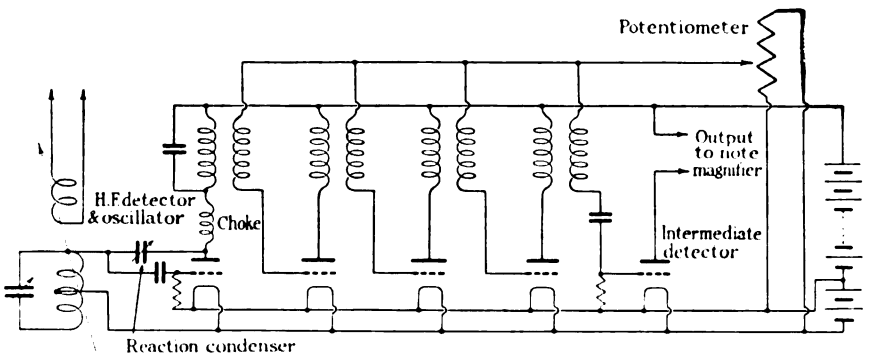

Fig. 15.—Circuit diagram of direction-finder supersonic amplifier.

The transformer and its mounting are shown in Fig. 16(*b*). The primary and secondary windings are separated by an "electrostatic" screen E which serves also as a mechanical support for the former. The screen consists of a cylindrical cage (8 in. high × 4 in. dia.) of fine vertical wires spaced $\frac{1}{8}$ in. apart. The method of winding the cage is perhaps of some interest. At the top the wires are supported by an ebonite disc which is carried on three brass legs fixed at their lower end to a brass ring to which the lower ends of the cage wires are all commoned. This ring is screwed to an 8-in. square brass baseplate (6) which carries the structure for supporting the 10-ft. horizontal arm. This brass plate is fixed to a brass tube (9) which forms the bearing, rotating about a fixed inner brass tube (10). This tube (10) has the down-leads spaced inside it and carries the transformer secondary at its top end, while its lower

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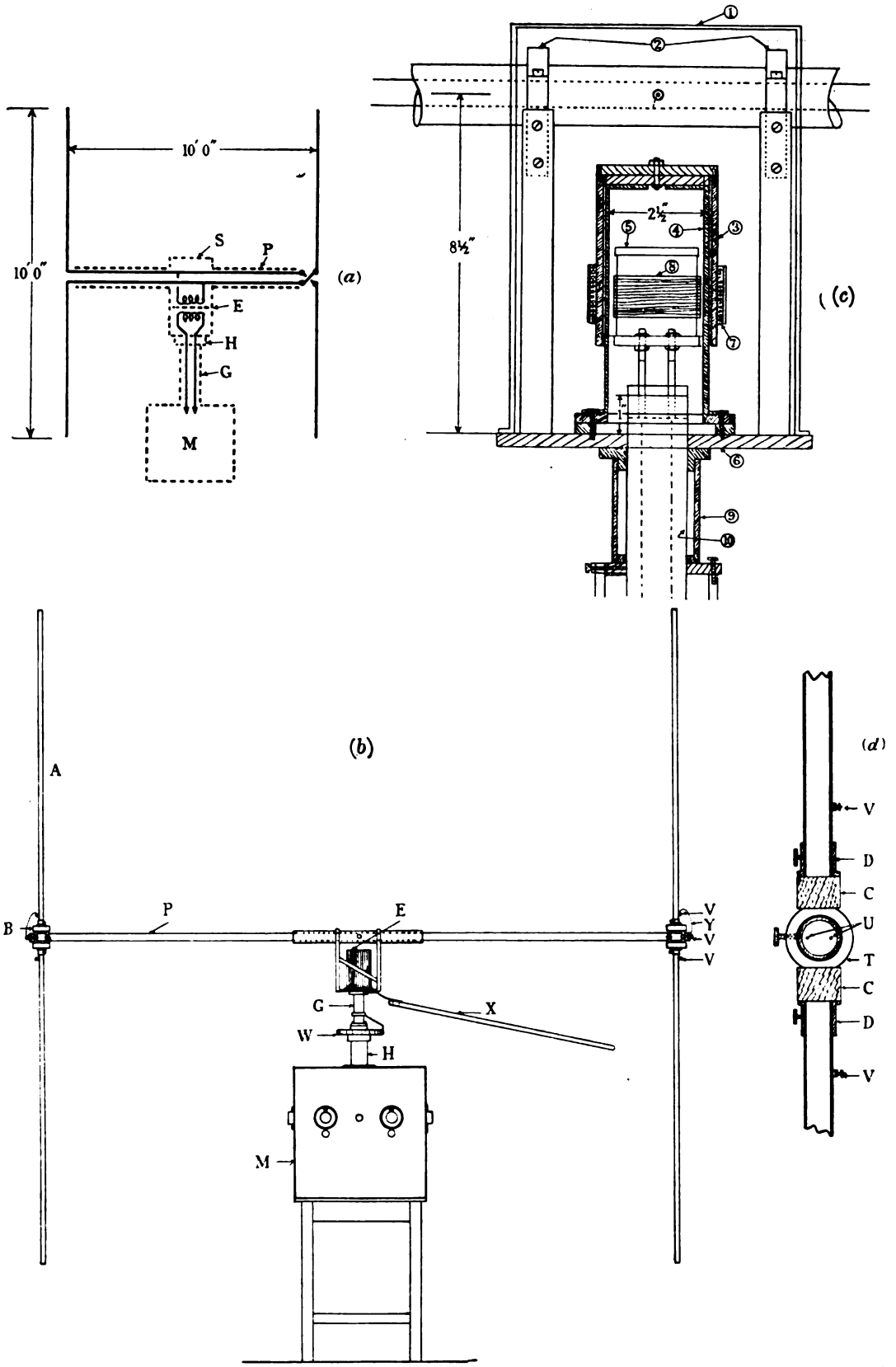

FIG. 16.—Rotating Adcock direction-finder. (*a*) Circuit and screening arrangements; (*b*) Mechanical construction; (*c*) Details of centre transformer; (*d*) Details of aerial junction-pieces.

A = aluminium rod aerial. B = junction pieces. C = ebonite spacing block. D = brass socket. E = electrostatic screen. G = brass tube with down leads. H = bearing. M = metal-lined box with amplifiers, etc. P = metal pipe. S = screen. T = brass ring with clamping screw. U = horizontal leads. V = terminals. W = scale and pointer. X = handle (projecting at right angles to horizontal arm). Y = connecting leads. (1) metal screening cover. (2) supports for horizontal pipe. (3) ebonite former for primary. (4) electrostatic screen. (5) ebonite former for secondary. (6) brass baseplate. (7) primary winding. (8) secondary winding. (9) outer rotating tube. (10) inner fixed tube.

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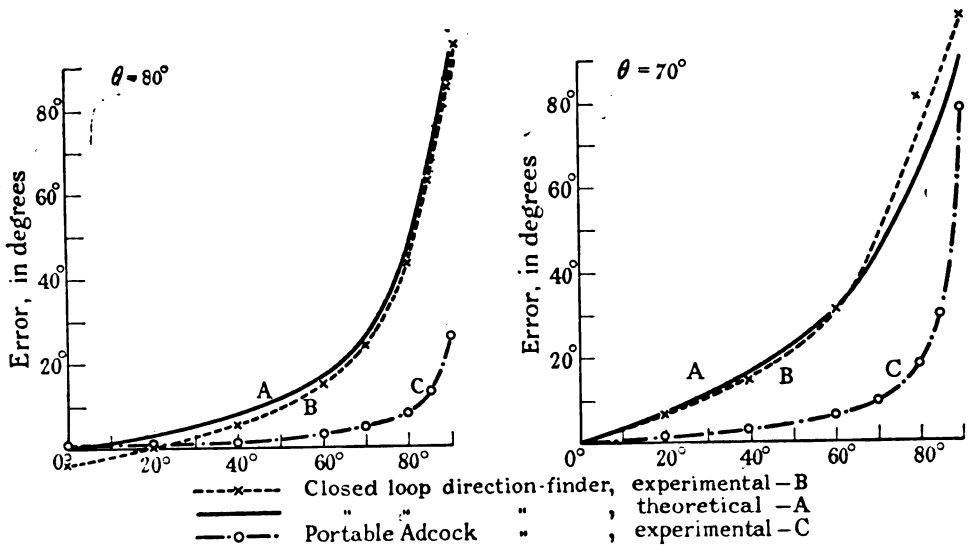
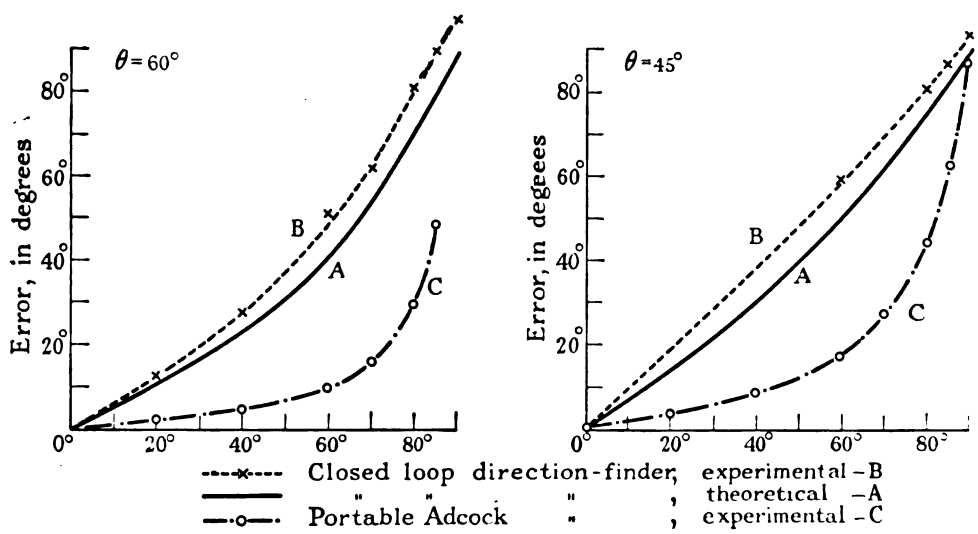
end projects downwards into the top of the amplifier and tuner box. The secondary coil (8) consists of a few turns wound on a 2-in. dia. ebonite former (5) which plugs into valve sockets on the top of the fixed tube (10). The primary coil (7) is wound on a 4-in. dia. former (3)

The aerial rods fit into brass sockets screwed to ebonite blocks mounted on brass collars sliding over the ends of the horizontal arm. The aerial rods carry terminals at their lower ends, and these are connected to the horizontal leads.



FIGS. 17 and 18.—"Aeroplane effect" direction-finding tests. $\lambda = 21$ m.



FIGS. 19 and 20.—"Aeroplane effect" direction-finding tests. $\lambda = 21$ m.

which slides over the screen (4). Thus the primary coil rotates with the aerial system but the secondary remains stationary, and since the coils are concentric there is no change in the mutual inductance. In this way direct mechanical connection between the rotating and fixed portions of the system is avoided.

The amplifier, its box, scale and scale mounting (B in Fig. 14) are actually the same as used for the screened direction-finding loop already described, though in this case the brass cylinder B no longer acts as a bearing but as a mounting for the scale and a support for the tube (9) which is locked inside it.

(3) Description of Tests.

The two types of direction-finders which have been described above have been subjected to a series of tests with the object of ascertaining their performance under varying circumstances. In particular, an attempt has been made to calibrate these instruments with respect to their behaviour when under the influence of downcoming waves of abnormal polarization.

As in the case of the medium-wave Adcock direction-finder described in Part 1, these latter tests are of two kinds:—(i) Tests with natural downcoming waves by

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observations on commercial transmitting stations at varying ranges and times of day, and (ii) tests with artificial downcoming waves provided by an elevated local transmitter—in this case not on a kite but fixed on the top of a 40-ft. telegraph pole.

be drawn with regard to the nature of the propagation of short waves.

Tests (*a*) (i) and (*b*) were made with the loop direction-finder only but they may be assumed to hold good for both instruments, for the two were frequently compared

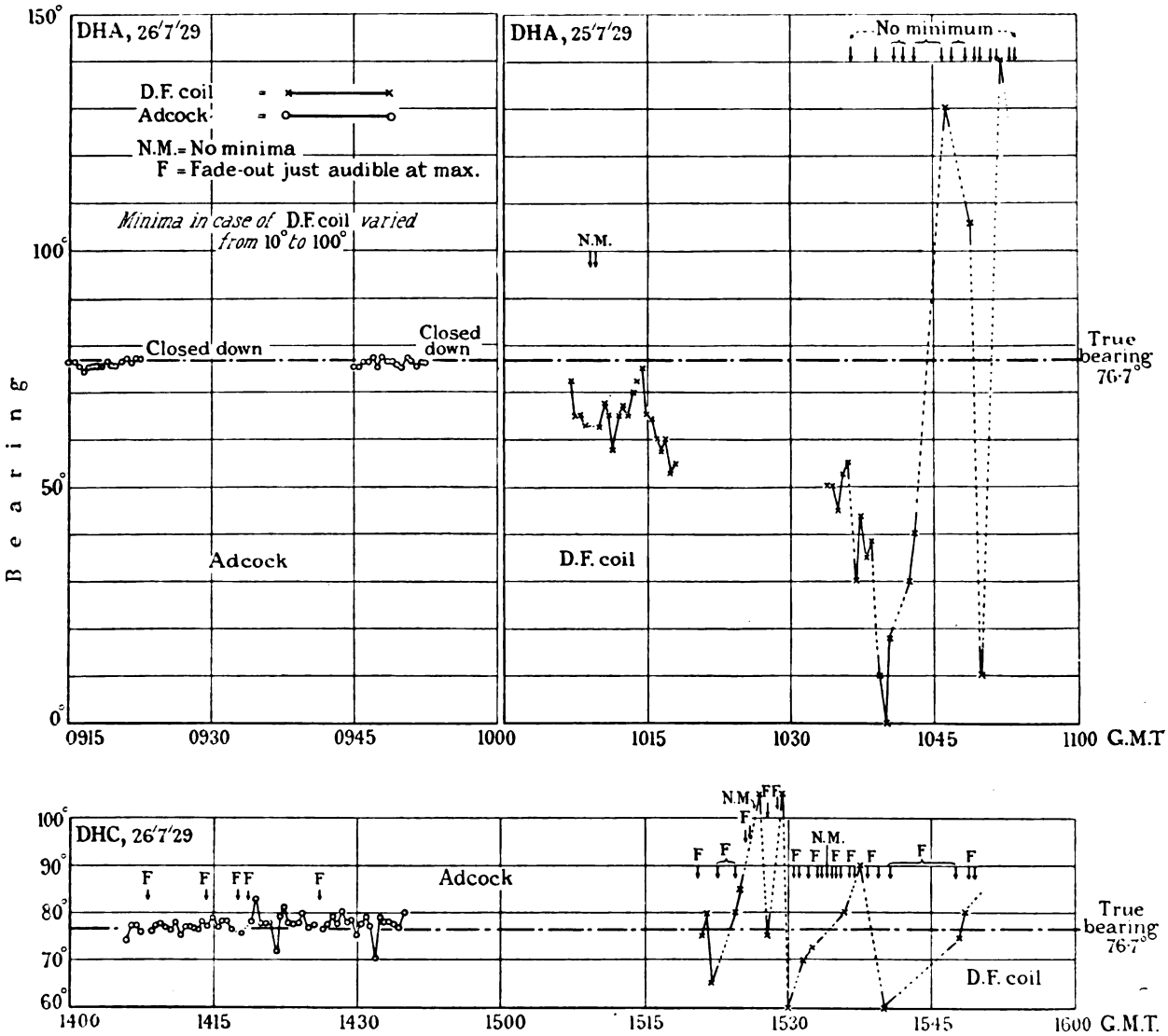

FIG. 21.—Comparisons of bearings on short waves taken by portable direction-finding coil and portable Adcock coil. Transmitters: Nauen [DHA (26·2 m) and DHC (22·7 m)].

The tests are divided into the following sections:—

(*a*) Tests with local transmitter.
  (i) Transmitter on ground.
  (ii) Transmitter elevated 40 ft. above ground.
(*b*) Effect of immediate surroundings.
(*c*) Tests up to limit of ground-ray range.
(*d*) Long-range tests.

Test (*a*)(ii) gives the inherent liability of the direction-finding sets to instrumental error, while the remainder give practical information about the conditions under which the sets can be effectively used. A consideration of the two tests together enables some conclusions to at close range from ground transmitters and found to behave similarly under these conditions.

(*a*) *Local Transmitter Tests.*

(i) With a low-power transmitter situated in the middle of a large, flat, open space and at a distance of 100 yards from the direction-finding set, the maximum error was 2·0° on all wavelengths from 16 to 40 m.

(ii) An important test carried out with the local transmitter was that of working with the transmitter in the air in order to provide a source of downcoming waves.

The transmitting aerial consisted of an aluminium

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rod 15 ft. long fed at the centre from a simple form of valve oscillator. This aerial was pivoted on a horizontal axis on the top of a telegraph pole 40 ft. above the ground. Directional observations were made on the ground at different distances from the foot of the pole with both types of direction-finding apparatus, in order to ascertain the error in observed bearing produced by the horizontally polarized component of the downcoming waves. The distances between transmitter and receiver were altered in successive stages to correspond to angles of incidence of the downcoming wave of 80°, 70°, 60° and 45° respectively, and at each site a series of observations was made for different inclinations of the transmitting aerial varying from 0° (vertical) to 90° (horizontal), thus providing waves of varying polarization.

The error ($\mathcal{E}$) which should theoretically be produced on the loop direction-finder in any given case is given by a simple calculation as

$$\tan \mathcal{E} = \tan \alpha \,.\, \cos \theta \,.\, \frac{1 + \rho_h}{1 + \rho_v} \quad . \quad . \quad . \quad (1)$$

where $\mathcal{E}$ = observed error,
θ = angle of inclination to vertical of transmitting aerial (polarization angle),
α = angle of incidence of downcoming wave, and
ρ_h/ρ_v = reflection ratio of ground to horizontally and vertically polarized waves respectively.

The results of the tests are summed up in the set of curves given in Figs. 17 to 20. Each figure is for a different angle of incidence (θ), and the curves each show the direction-finding error with respect to α, (*a*) being the theoretical error curve for the loop, (*b*) the experimental error curve for the loop obtained from (1), and (*c*) the experimental curve for the Adcock direction-finder.

The theoretical loop curves agree very closely with the practical loop curves, and in all cases the Adcock errors are seen to be notably less than the loop errors. The extent of the superiority of this type of Adcock direction-finder over the loop type for observation on downcoming waves is exactly represented by these curves and cannot be expressed in any better way. It is of interest, however, to write down one or two examples. Thus, for a wave arriving at an angle of incidence of $\theta = 80°$ (10° to horizontal) and polarized with its electric force at $\alpha = 45°$ from the vertical, the error on the loop is 10°, while on the Adcock it is less than 2°. Again, for $\theta = 45°$, $\alpha = 60°$, the error on the loop is 60° and that on the Adcock 18°.

The cause for the existence of the residual error revealed in Figs. 17–20 is probably the asymmetry introduced by the earth's surface. A successful attempt was made to reduce these residual errors by shortening the lower limbs of the aerials. The curve obtained by cutting 2 inches off these aerials is shown in Fig. 22 for one angle of incidence. Tests with other angles of incidence, however, still showed an improvement, but not so great.

(*b*) *Effect of Immediate Surroundings.*—These tests were undertaken mainly to provide information affecting choice of site for short-wave working. Observations were made to determine the effect of local contour variations, villages, individual houses, boundary walls, etc. The general conclusions arrived at were that, provided the same precautions as to choice of site were taken that are taken with longer-wave direction-finding apparatus, the same general accuracy with respect to such local effects will be obtained as is obtained with the longer waves.

A few examples are given below to illustrate the nature of these tests.

(i) Observations were made in a large field at a distance of about one mile from the Leafield transmitter. The field was approximately level and had an area of about 100 acres. Eighteen different positions were investigated in this field on a wavelength of 17 m. Eleven showed no error exceeding 1°, and individual readings could be taken to within ½°. Of the remaining seven for which the error was greater than this, one was within 10 yards of a cottage and had an error of 7°, and six were within 200 yards of a small village which lay between them and the transmitter. The errors

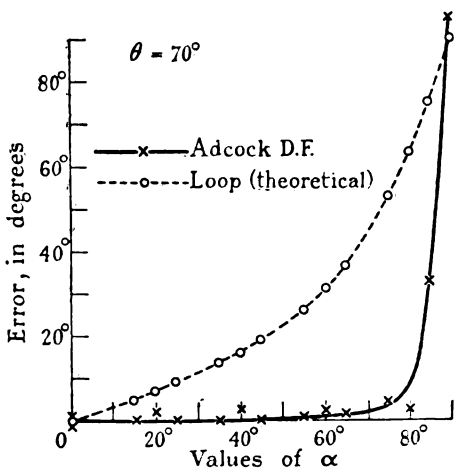

FIG. 22.—Tests on rotating Adcock direction-finder. Balanced aerials:—top half 5 ft., bottom half 4 ft. 10 in.

increased steadily as this village was approached, reaching 2½° at a distance of 100 yards, while at closer range the minima became too flat to allow observations to be made. Although the field was within a mile of the transmitter, the intervening ground was by no means level. The waves in their passage had to pass over a deep fold in the hills which cut obliquely across their path. This feature, as the general freedom from error indicates, did not cause their course to deviate to any appreciable extent.

(ii) Observations were made on the error produced by a single tree 70 ft. high. At a distance of 10 ft. from the tree the maximum error was 40°; as the distance was increased this rapidly fell off until at 30 ft. it was 10° and at 100 ft. it was negligible.

(iii) Observations made near a stone boundary wall about 3 ft. high showed no appreciable error even when the distance away was only a few feet.

(*c*) *Ground-Ray Tests.*—The two types of apparatus (loop and Adcock pair) have been thoroughly tested on a wavelength of 24 m over a range of from 5 to 150 miles, i.e. up to the point where the ground ray disappears.

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TABLE 3.—*Short-Wave Direction-Finding Observations on Adcock and Coil Portable Direction-Finding Apparatus.*
Transmitter at Chelmsford (5SW, $\lambda = 24$ m)

No.	Date	Time, G.M.T.	Place	Nature of site	Distance	True bearing	Error (mean of several observations)		Remarks
							Coil	Adcock	
					Miles				
1	5.6.28	Mid-day	Terling, Essex	Many local trees. Site otherwise good	6	223°	− 0·3°	− 0·3°	—
2	5.6.28	Evening	Bee's Farm, near Halstead	Bad. On grass space in centre of cross-roads	15	208°	+ 7°	+ 6°	—
3	6.6.28	Mid-day	2 miles N. of Sudbury on side of Sudbury–Melford road	Bad. D.F. only 60 yards from wires and very small field	24	205°	− 8°	− 9°	Marked fading
4	6.6.28	Evening	Thorpe Morieux, Suffolk	Good, open site in sloping field (5° to S.W.)	32	209°	+ 3·5°	+ 1·4°	Fading very strong, very difficult to take readings
5	7.6.28	Mid-day	Ixworth, Suffolk	By roadside after heavy rain. Bad site	43	202·5°	− 1·2°	− 2·5°	Fading very slight, but flat on coil and difficult to get bearing
6	7.6.28	Evening	New Buckenham, Norfolk	Very good site on large flat common	58	206·5°	—	—	"Night effect" too bad to take bearing
7	8.6.28	Mid-day	New Buckenham, Norfolk	Very good site on large flat common	58	206·5°	—	− 1·5°	Considerable fading
8	1.8.28	Mid-day	Wansford, Huntingdonshire	Good, open site 200 yards from North Road	73	148·5°	− 3·5°	− 3·5°	Signals rather faint and minimum flat and ill-defined
9	2.8.28	Mid-day	Waddington	Very good site on plateau	110	158°	− 3·0°	− 8°	Minimum very flat on both systems and all but impossible to obtain
10	2.8.28	Evening	Waddington	Very good site on plateau	110	158°	—	—	No minimum obtainable on either system
11	3.8.28	Mid-day	5 miles N. of Doncaster on Selby Road	Good site, though well-wooded district	150	152°	—	—	No minimum obtainable on either system

The plan of these tests was to limit the observations to one transmitting station only (Chelmsford 5SW, $\lambda = 24$ m), and to make these observations as nearly as possible simultaneously on both systems and at gradually increasing distances from the transmitter in approximately the same direction. The observations were made in a northerly and north-westerly direction from Chelmsford, the maximum and minimum distances from the transmitter being respectively 150 and 6½ miles. Observations were made at some sites at mid-day and at others in the evening about an hour before sunset, while again in some cases the same site was investigated in the morning and in the evening. At night it becomes difficult to take readings even at a distance of 30 miles, owing to fading effects. In the daytime 70 miles appears to be the limit, though at 110 miles it was possible to get a rough bearing. The results are shown in Table 3. The maximum fixed error observed, where observations were possible, was 9°. This was probably an error local to the site, which in this case was a bad one. On the whole, the results within the range of possible bearings stated are good and compare favourably with those obtained on longer waves under the same conditions.

It was noticeable in contrast to medium- and long-wave direction-finding experience that there were no "variable errors." When fading and the ordinary symptom known as "night effect" occurred, the changes were too rapid to enable bearings to be taken during a swing.

(*d*) *Long-Range Tests.*—To give some idea of the working of the apparatus at points beyond the "skip distance" a short series of observations were taken at Slough on the Nauen stations AGB. The loop receiver gave very large errors, the minimum shifting to all parts of the scale indifferently. At other times there was a faint indication that the very flat minimum oscillated about the true bearing. On the Adcock, however, the minimum, though shifting rapidly over a small arc, which would sometimes be as great as 10°, could always be found to average within 2° or 3° of the true bearing. The curves in Fig. 21 show the nature of the results obtained.

(4) Conclusions.

(i) The results obtained on the long-range tests (*d*), taken in conjunction with the local tests with the elevated transmitter (*a*) and those within the skip distance (*c*) (see Table 1 and Figs. 17 to 20), confirm the conclusions arrived at by Eckersley* from experiments of a somewhat similar nature. The variable errors and flatness of minima observed at distances exceeding 20 miles (and less than, say, 500 miles) are undoubtedly due to downcoming waves. The fact that at such ranges these effects were of equal extent on both loop and Adcock direction-finders is a clear indication that these downcoming waves are scattered or laterally deviated in the course of their reflection from above, for had the waves remained undeviated laterally the Adcock direction-finder would have shown less errors than the loop, as is clearly illustrated by the curves of Figs. 17 to 20.

* *Loc. cit.*

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(ii) With either of the instruments described, direction-finding is possible on a wavelength of 24 m up to a distance of 70 miles in the daytime and 30 miles at night (these results applying to summer season only). Up to a distance of 20 miles the bearings are sharp and accurate to within 2°, provided a good open site is found (say 100 yards clear of obstacles in all directions). At distances greater than 20 miles minima become flatter and flatter until they become too flat to read, on both types of direction-finder. At much greater distances, say more than 500 miles, the Adcock pair gives fair indications of bearings of errors not greater than 5°, while the loop direction-finder is practically useless on account of flatness of minima and extent of bearing variation.

(iii) The Adcock direction-finder, though not completely immune, is much less subject to errors due to downcoming waves (provided, of course, they are not scattered) than the loop apparatus, to an extent which is best expressed by the curves of Figs. 17 to 20, and it is for this reason more suited for use with aircraft or for long-distance working from ground stations. At close ranges to ground stations (say within 20 miles), where the downcoming radiation is weak, scattered or very steeply inclined, since there is nothing to choose in accuracy of results the loop direction-finder is to be preferred, being more compact, robust and portable.

This work was carried out as part of the programme of the Radio Research Board and is published by permission of the Department of Scientific and Industrial Research. The author wishes to acknowledge his indebtedness to the staff of the Radio Research Station at Slough and the assistance of many of his colleagues in the work. He would also like to thank Mr. Watson Watt for valuable suggestions relating to the presentation of the results. Much of the observational work was carried out by Mr. Chapman and Mr. Munro, and Mr. Hayman assisted in some of the experiments.

Discussion before the Wireless Section, 2nd April, 1930.

Mr. N. E. Davis: The author has obtained an excellent result in a difficult and patient research toward the elimination of "night effect" in radio direction-finding—a problem to which the research department of the Marconi Co. also continues to devote considerable time. The full practical realization of the Adcock principle for eliminating "night effect" depends almost entirely on providing a non-radiating link between the vertical aerials and the radiogoniometer. The problem is to devise a practical method of transferring energy from the vertical aerials to a receiver situated at ground-level, in such a manner that only the normally polarized components of the wave give directional indication. I should like to mention the method which we have developed for providing a solution of this problem, and the practical results so far obtained. The Marconi adaptation of the Adcock principle for indicating the incoming direction of short-wave radiation relies firstly upon effective screening of the horizontal members, and secondly upon nullification, so far as this is possible, of re-radiation from the screen to the active vertical aerials. The success obtained led us to believe that similar methods applied to the Adcock principle on the medium and longer wavelengths would probably provide a solution to the elimination of night errors in practical direction-finding. Owing to the increasing importance of producing reliable direction-finders as night navigation aids for aircraft, it was decided to make initial experiments on wavelengths of the order of 900 to 1 000 metres. For this experiment, four vertical aerials 70 ft. high were erected with a spacing of 100 yards. This represents a spacing of approximately $\frac{1}{10}$ wavelength. Each of the aerials was terminated in a screening box situated about 5 ft. above the ground and in the same vertical line as the aerial. A single conductor completely shielded in copper tubing and carried about 2 ft. above the ground constituted the horizontal lead-in arrangement from each aerial. This was led to a carefully shielded radiogoniometer in the hut. The ends and centre of the horizontal shields were earthed. Observations on this system over a period have shown that with this adaptation of the Adcock principle it is possible to guarantee an accuracy of 2° to 3° under the worst conditions of "night effect," on a wavelength of the order of 1 000 m. Our experimental apparatus as set up has a direction-finding range of the same order as that of the ordinary loop direction-finder, and it will fulfil commercial or service requirements. We are not yet certain what wave-range can be obtained on a given installation, when keeping within the above-mentioned limits of accuracy. The research department of the Marconi Co. thought that it might add to the interest of the meeting and to the discussion, as well as to the scope of the paper, if I were to show a few lantern slides of the arrangement of our experimental station of medium wavelength. [Mr. Davis here exhibited four slides illustrative of the Marconi Co.'s station.] Tests on wavelengths of the order of 450 to 500 m give every hope of an all-round direction-finding accuracy closely approaching that obtained on the longer wavelengths.

Mr. S. B. Smith: The medium-wave apparatus is of particular interest to most of us, as it is on these waves that radio navigation can be successfully carried out. I should like to ask the author his opinion regarding a suitable wave-range for his apparatus; no doubt commercially it would be ideal to operate the equipment on all usable wavelengths, but does a 2 : 1 wave-range *without rebalancing* represent what he considers to be practical? Eckersley's work on the short-wave Adcock system indicates that it is possible to obtain accurate indications though they may bear little relation to the geographical position of the transmitter, provided the feeder system connecting the aerials to the radiogoniometer is really efficiently screened, and it does seem that, if a solution on these lines can be obtained, balancing schemes, which always present practical operating difficulties, will not be required. The tuned Bellini-Tosi system can be made to give very accurate results in daytime, but the need for frequent and accurate rebalancing made the development of the so-called "aperiodic

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system" a real necessity. Recently one has read a good deal about directive beacon transmitters of various types, and of their utility as an aid to air navigation; it does seem to me that if night ranges greater than 100 miles overland are required, then the Adcock method in some form must be a natural development, and the present loop methods will become obsolete. The problem of producing a radio beacon free from night effect may become even more difficult than that of the direction-finder, since it appears likely that the equi-signal beacon method of transmission will not prove to be free from night effect when applied to an Adcock aerial system, unless very careful symmetry be maintained between each aerial and the feeder system. In plotting the deviation curves on the Adcock and loop arrangements I notice that the author has omitted to record any observations on the Adcock aerials when quite large variations are present on the loop; why not give definite observations in such cases? It has been our custom in plotting swings greater than 90° to reverse the sign of the observations. The chain-coupled Adcock direction-finder shown in Fig. 12 suggests another method which in practice may be even more difficult to realize; this consists in coupling each aerial to a rectifier, radiating a local carrier from a central aerial, and passing a beat tone over four lines to a central hut. It should then be possible by this frequency change to construct a line of such constants that at the fundamental frequency it is non-radiating, while the changed frequency can be arranged to reproduce the desired results at the radiogoniometer, possibly at low frequencies. We tried this scheme on a wavelength of 15 000 m in 1923, but the test failed owing to the small receiving power of the aerial. The spacing between the aerials was only 1/200th of a wavelength. In comparing night variations on direction-finders spaced about 2 wavelengths apart it has been found that the deviations are far from synchronous, and although the variations do usually occur on both aerials a very small interval may pass before the variation will be reproduced on the second direction-finder.

Mr. C. E. Horton: The main results of the author's work are entirely in agreement with my own experience. With regard to the last sentence on page 173, to the effect that the range of the installation with standard forms of amplifier was not noticeably different from that of the Bellini-Tosi system, I think it might legitimately be inferred from this that the Adcock is just as sensitive as the Bellini-Tosi direction-finder on medium wavelengths. A Bellini-Tosi system with square loops of 20-ft. side constitutes a satisfactory and sensitive receiver for wavelengths between 200 and 20 000 metres. Would an Adcock direction-finder be equally sensitive under the same conditions of space and wavelength range? I think that its sensitivity would be less, because the impedance of the aerial system of the Adcock apparatus on long waves compares unfavourably with that of the Bellini-Tosi equipment. The aerials used in the author's experiments are very large—80 ft. high and with 50-ft. spacing. A Bellini-Tosi system of corresponding size would give a large response. With a medium-wave Adcock system of the size of an ordinary frame coil it would be very difficult to get sufficient signal strength. With regard to residual errors in the author's system, it seems to me to be unfortunate that he did not use another system which has been employed in the Navy and which, while not theoretically faultless, gives very good results—possibly better than those in the paper. I refer to the method of eliminating all horizontal pick-up by using lead-sheathed cables. All horizontal leads take the form of paper-insulated cables several feet underground, as used by Mr. Eckersley in his short-wave work. For ground of ordinary conductivity this is a simple and satisfactory way of arranging Adcock aerials, and it would be of interest if the author were to continue his work by making a direct comparison between this method and his own. I am particularly interested in direction-finding in ships, where I found that the Adcock system was difficult both to install and to work. It is assumed in the theory that two aerials of equal height have equal e.m.f.'s induced in them, and serious errors will result if equal pick-up is not obtained on all the aerials. Two vertical aerials of exactly equal dimensions installed in a ship are generally found to differ perceptibly in receiving power, and it is only by careful adjustment, usually varying with the frequency, that exact equality of induced e.m.f.'s can be obtained. I refer chiefly to battleships, where the congestion of the metal work on the upper deck considerably affects reception. It is possible that equal difficulty is not experienced in merchant ships. On page 186 the author refers to the necessity of a good open site (say 100 yards clear of obstacles in all directions). Such conditions are not found on a battleship, and remembering that an error of 2½° was obtained in a bearing owing to a village 100 yards distant, the effect of a funnel a few feet away and the difficulties associated with short-wave direction-finding in ships can be appreciated. Oddly enough, better results have been obtained on very short waves of the order of 12 m than on waves of the order of 30 or 40 m.

Mr. J. F. Herd: The difference in the behaviour of the closed-coil receiver and that of the Adcock in the presence of abnormally polarized downcoming waves is beautifully demonstrated by the small elevated transmitter (with which the author's curves of Figs. 17 to 20 were obtained). While the transmitter is giving normal polarization, the closed-coil receiver gives good and accurate minima. When the transmitter is tilted so as to give abnormal polarization the closed-coil receiver gives a minimum which is visibly inaccurate. This is particularly obvious if the coil is kept set at the minimum for the transmitter and the transmitter is then swung. Very considerable re-orientation of the coil is necessary to give even the inaccurate minimum. Repeating this experiment with the Adcock aerial, while the polarization is normal the directional indication is exactly similar to that obtained with the closed-coil receiver. When, however, the transmitter is tilted by the same amount as in the coil experiment, the minimum indication does not wander. This is again particularly obvious if the Adcock aerial is kept set at the transmitter minimum, and the transmitter then tilted, when no change of minimum is discernible. With reference to the medium-wave apparatus, it would be interesting to have a comparison between the results obtained by Dr.

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Smith-Rose and the author when working in their original elevated hut, and those given by the present Adcock system (of Figs. 1 and 2) working at ground-level.

Dr. R. L. Smith-Rose: When Mr. Barfield and I wrote our paper* some three years ago we left the Adcock direction-finder literally in the air. The apparatus was suspended between the poles carrying the aerial system. The work of bringing the system down to earth, which the present author has carried out, may seem to be a simple operation so far as the hut is concerned, but the difficulties encountered in connection with the rest of the apparatus can only be fully appreciated by those who have had considerable experience of this class of research work. On looking through Mr. Adcock's patent recently I was surprised to find how completely its originator had appreciated the nature of the problem even so early as 1916. He has described in full a number of methods whereby the difficulties due to the downcoming horizontally polarized electric waves could be overcome. It appears, however, that his conception of this matter was rather too early for it to be fully appreciated by those responsible for the application of direction-finding to navigation. In the present paper the author has shown very clearly, in such diagrams as Figs. 17 and 18, the possibility of considerable errors, of 20° or more, being obtainable with the ordinary closed-loop direction-finder when used in making observations upon the transmissions from aircraft subtending an angle of elevation at the ground of the order of 10°. As the diagrams in the paper show, these errors can be reduced to about one-tenth of their magnitude by the use of the Adcock aerial system. I would suggest that it is important that the development of the apparatus in the direction of its practical application should be proceeded with as rapidly as possible, in order that it may be retained as an entirely British invention. I believe that no other country is at present using it for practical direction-finding, although the system is being used in America for studying the propagation of waves. In connection with the theoretical part of the first section of the paper given on pages 176 to 178, I should like to ask the author whether he has not taken into account twice some of the e.m.f.'s induced in the aerial circuit. It appears to me that the equation in col. 2 of page 176 contains an expression of the current as derived from the electric force acting on the aerial system, plus the same quantity as derived from the magnetic force acting on the system. I would suggest that it is an axiom that, in dealing with electromagnetic waves and their effect upon aerials and loops, one may consider the effect of either the electric field or the magnetic field. By suitably manipulating the necessary equations one should obtain the same result from either method. For example, in the case of the induction of an e.m.f. in a closed loop by a stream of electromagnetic waves, one may take either the line integral of the electric force round the perimeter of the loop or the surface integral of the magnetic force intercepting the loop at right angles to its plane. Since the magnetic force is merely derived from the motion of the electric force the same result for the induced e.m.f. will be obtained for either case. I am aware that in certain cases, such as that of the reflection of waves from a conducting surface, it is possible to obtain a resultant magnetic component which is not equal to the resultant electric component in a certain direction, and the measurement of the ratio of the intensities of these components is used as the basis of the method of measuring the angle of incidence of waves upon a conducting surface. In this case, however, we are dealing with resultant fields in two directions, and a detailed consideration of the directions of the fields and of their motions will show that the conclusions stated above are not in any way invalidated. A portion of the analysis given on page 177 is somewhat difficult to follow, apparently on account of one or two misprints and omissions in the nomenclature, which I have already pointed out privately to the author. I have always found the aerial system drawn in Fig. 9(*a*) a little difficult to follow, especially when one is attempting to visualize the conditions which are necessary for compensation of the e.m.f.'s induced in the horizontal portions. If, however, the aerial circuit is re-drawn as a simple impedance network, as indicated in Fig. A,

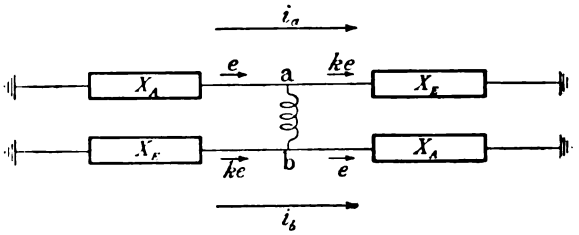

FIG. A.—Condition for zero current through "a b" is $X_A = X_E$.

it then immediately becomes apparent that in the presence of horizontal electric forces from the arriving waves the condition for obtaining low current through the field coil "a b" is that the reactance of the aerial must be equal to that of the earth connection beneath it, assuming that the four aerials and earth connections are identical in dimensions and configuration. This condition can be met either, as the author has stated, by introducing negative reactance into the earth lead or, alternatively, by introducing positive reactance into the aerial. Since in the case of the direction-finder it is desirable to maintain the natural wavelength of the aerial system well below that of the arriving waves, the alternative of introducing negative reactance into the earth lead is obviously the one to adopt. As the author has pointed out, a further condition to be met in order that a complete balance of the aerial system shall be obtained is that the current distribution in each aerial circuit shall be symmetrical about the centre point of the coil "a b." In col. 1 of page 177 the author states that this condition is obtained at the same time as that of the equality of reactance by the use of balancing capacity in the earth lead. As the discussion of this condition is dismissed in a few lines and the conclusion is not quite obvious, I should like to ask the author whether he can elaborate this point a little and give a more detailed explanation. In con-

* "The Cause and Elimination of Night Errors in Radio Direction-Finding," *Journal I.E.E.*, 1926, vol. 64, p. 831, also *Proceedings of the Wireless Section*, 1926, vol. 1, p. 115.

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sidering the possibilities of the future development of the Adcock system, I am not clear from the discussion given in the paper to what extent the systems indicated in Figs. 12 and 13 are superior to that of Fig. 11, which appears to me to offer the most hopeful promise of success in the future. The author admits on page 179 that the arrangement shown in Fig. 11 is subject to only one type of error, viz. that due to the effect of the electric coupling between the horizontal conductors and the vertical aerial. I would suggest that there is also the possibility of the induction of an e.m.f. directly into the loop enclosed by the horizontal conductors. I think that both these errors could be reduced to zero by a suitable arrangement of the horizontal conductors between the two vertical aerials. In the second part of the paper the author has shown that with the aid of modern receiving amplifiers it is possible to use direction-finders of either the closed-loop type or the Adcock type on wavelengths from 12 to 60 m with similar success, as far as the actual instrument is concerned, to that obtainable on wavelengths of 300 to 1 000 m. The author has displayed considerable ingenuity in this short-wave direction-finder in arranging the coupling coil between the rotating aerial system and the secondary circuit with no mechanical connection between them. The latter portion of the paper describing the results of observations on short-wave transmissions is important not only in showing the limitations of any system of direction-finding on these wavelengths, but also in demonstrating the utility of the two types of direction-finder in studying the propagation of short waves.

Lieut.-Col. H. P. T. Lefroy: In communication by line, the " rotation " of the go-and-return conductors, to avoid effective radiation from, or absorption by, those conductors, is a well-known and very useful arrangement; the pitch of such mutual twisting of the two conductors about each other is relatively long, and the frequencies dealt with are of the acoustic order. In 1921 I investigated the effect of such mutual twisting for radio frequencies, and found that the shorter the pitch the more complete was the absence of radiation from, or absorption by, a pair of go-and-return conductors. The higher the frequency dealt with, the shorter should be the pitch. I tried twists up to 10 per inch, using silk-covered wires of suitable gauge. The effect of such intensive twisting is greatly to increase the mutual capacity and reduce the inductance of the pair; the effective resistance is also increased. When such a sharply-twisted pair, joined at one end, was wound to form the secondary of a radio-frequency transformer, no sound could be heard in a telephone fitted, in series with a rectifier, across the other end of the pair, when the coil was suitably coupled to a primary coil carrying fairly strong damped oscillations. It seems possible that all undesired " pick-up effect " might be eliminated, in the Adcock direction-finder for short waves, if such an intensively-twisted pair were fitted to act as the horizontal leads, with the receiving coil connected, between the centre points of the two wires forming the pair, by a small fixed condenser at each end of that coil.

Air-Commodore A. D. Warrington-Morris: I am glad that this paper on Adcock aerials has been read, because the subject has recently been arousing considerable interest from the points of view both of Service flying and of civil aviation. Three of the civil aviation companies—the French, German and Belgian—are starting night services to Croydon. We have three directional stations—at Croydon, Pulham and Lympne—which by day normally guide the movements of the international aircraft. They do it very satisfactorily, and in foggy weather their function is an important one. Now that night services are being started it is necessary to produce direction-finding apparatus which will be of use after dark. The results obtained with the Adcock aerial are strikingly good, but I am not clear as to what is meant by " standard deviation." The author states that it is obtained by taking the root-mean-square value of the variation of individual observations from their arithmetic mean: I should like to know how this compares with what we call the " average error " (plus or minus). The distance between the two vertical aerials shown in Fig. 1 is given as 50 ft. I understand that accuracy, as well as pick-up value, is affected by the distance apart of such vertical aerials, and I should like to ask the author whether it was purely with a view to economizing in space and the cost of masts, or whether it was due to the difficulty of balancing, that he only allowed 50 ft. between the two vertical aerials. In connection with this last difficulty, is it necessary to re-balance the system after a change of wavelength? I recently saw a demonstration at Chelmsford of a Marconi-Adcock arrangement in which four 70-ft. masts were employed, and the leads-in were thoroughly screened. This seemed to me to be a simple and efficient method of attaining the object in view, i.e. the elimination of the effect of the horizontal aerial. A number of bearings on Hilversum were taken at night simultaneously with this equipment and an ordinary Bellini system. Whereas the maximum error with the Adcock system was about 2°, the Bellini apparatus gave errors of 20° or more. With regard to the author's conclusions as to short-wave direction-finding, he deals a blow to the installation of short-wave apparatus in aircraft when he indicates that it will never be a practical proposition, on wavelengths of 12 to 60 m, to take direction-finding bearings of a transmitting station at ranges exceeding 20 miles. I should like to ask him whether there is any hope of success on rather longer waves of from 60 to 100 m, provided that the power of the transmitting set is considerably increased. A range of about 200 miles is required.

Mr. T. L. Eckersley: I am glad that the author has thoroughly tested-out the behaviour of the Adcock aerial under what may be called horizontal polarization conditions. Although my main interest at present lies with short waves, direction-finding results on the long-wave band which might affect this long-outstanding problem are of great interest to me. The author has produced at least one method of obtaining bearings free from night error. There are, I think, other systems capable of achieving this result, and it is merely a matter of practical handling to determine which is likely to be the best. Regarding the circuit shown by the author, in which horizontal lead-in is balanced by horizontal lead-out action and then balanced by a condenser to

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earth, I would mention that about 1917 I made a similar suggestion in a report to the War Office. It is easy, however, to make a suggestion: it is to the application of the suggestion to a practical system that the real credit is due. With regard to short waves, the author has done much of the work which I had left to future developments. I am glad to find that he affirms in a general way the results which I obtained earlier with rather a different type of Adcock apparatus, but one which appears to have functioned very similarly to the Adcock aerial which he himself has used. The short-wave frame apparatus with its screened loop seems to be a good arrangement as it almost entirely eliminates the vertical effect. By means of careful balancing, however, it is possible with a frame to get almost perfect minima on short waves if everything such as batteries, receiver, etc., is shielded. At present we are using a receiver in which the batteries are external with a short-wave direction-finder, and it is only with a 2-ft. frame that we find it possible to anchor the leads outside in such a way that the amount of energy received on battery leads, etc., is negligible compared with the amount received on the frame. Such an arrangement dispenses with the necessity of having batteries inside a portable box, and to a certain extent this makes it easier to work. With regard to the Adcock arrangement for short waves, this seems to be an alternative method of obtaining signals free from " night effect," and, as the author points out, it is rather difficult at first to see why the balance should not be perfect. I agree with him that this is not possible because of the fact that the lower limbs of the vertical part are not exactly symmetrical with the upper part. In connection with the long-distance test, I should like to ask whether the author made any measurements on distant stations, such as New York, so as to gain some idea of the accuracy obtainable on these. On our system I have noticed a swinging of the bearing of about 3° or 4° on long-distance stations, for example Canada and New York, and I am not sure whether to attribute it to a real wobble of the bearing or to horizontal polarization effect. I think that far better balancing than can be obtained either with the author's or my own system is necessary in order to make certain of this. The question would be important from the point of view of beam transmission should a wobble of 3° or 4° be shown to exist. I think that the author has confirmed the presence of scattering and, perhaps in response to this, one may say that there is little hope of getting reliable direction-finding results in the " skip distance." As a matter of fact, errors up to 90° may be obtained in the case of the beam transmitter, and one never knows where this may occur. So far as we know at present, it is impossible to be certain of the bearing within the " skip distance."

Mr. F. Adcock (*communicated*): In the course of the discussion on the paper a speaker referred to a direction-finder which appeared to be remarkably free from " night error." In that equipment the horizontal insulated leads were screened by placing them within metal tubes which were carefully sealed and earthed at several points. I am interested in this method of reducing or avoiding " night error," as at one time I experimented with a direction-finder in which the horizontal leads from the vertical aerials to the operator's hut consisted of insulated wires simply laid on the surface of the ground. With this improvised arrangement the errors in bearings taken on aircraft were about one-quarter as great as those procured by means of a closed-loop direction-finder. As it was impossible to make further experiments it remained doubtful how far the improvement in accuracy would be maintained for other wavelengths and for different conditions of the soil on which the direction-finder was erected. One would expect a reduction in " night error " when the horizontal leads of a direction-finder are placed near the ground-level, but the complete elimination of the error in all circumstances is unlikely by this method, owing to the poor conductivity of the soil. So long as the soil in the vicinity of the direction-finder is fairly uniform in conductivity the re-radiated energy takes the form of the usual reflected wave, which should not unduly disturb the bearings obtained by the installation. On the other hand, horizontal conductors situated near the direction-finder would distort the reflected wave. Expressed in another way, oscillations present in horizontal conductors or in the earthed screens surrounding horizontal conductors may, owing to re-radiation or to electric or magnetic coupling, give rise to undesired oscillations in the vertical aerials. Owing to the relatively poor conductivity of soil it is impossible to " anchor " the screening tubes by any ordinary system of earthing. For these reasons it appears that there is some risk of error in direction-finding installations where the horizontal leads are carried in long continuous screening tubes which are earthed at several points.

Mr. R. H. Barfield (*in reply*): I am interested to hear that Mr. Davis is adapting Eckersley's screened " U " Adcock to practical direction-finding, and to see that it is giving good results. Experiments which I have carried out since reading the paper show that for short waves (20 metres) the screen can have but a negligible effect on the performance of the system, whether earthed or insulated. The ordinary unscreened " U " Adcock is amenable to fairly simple calculation, and it appears that the horizontal pick-up may be of the same order as the vertical pick-up under some conditions. For this reason it was dismissed some time ago. Mr. Adcock gives a criticism of the "U" Adcock screened and unscreened, with which I am in substantial agreement. He gives a figure for the performance of the " U " Adcock which, though somewhat more favourable, is of the same order as that obtained by a calculation.

In reply to Mr. Smith, the wavelength range of the balanced Adcock system with a given balancing capacity has no upper limit other than that of amplifier sensitivity since, for waves long compared with that of the aerial, the impedances which have to be balanced are both purely capacitative and therefore change equally with frequency. The lower limit occurs when the working wavelength begins to approach that of the aerials. It should therefore be possible to work with very large wavelength ranges. For reasons connected with the residual effect referred to in the theoretical analysis, the balancing of the condenser is not critical, and the case of the tuned Bellini system is therefore not a parallel one. In a commercial design of the system they would probably

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take the form of fixed condensers in sealed weather-tight boxes and would never require any attention. The interruptions in the Adcock observations, referred to by Mr. Smith, were due to causes such as adjustment of apparatus, changing of batteries, etc.

In reply to Dr. Smith-Rose, in assuming a sine-law distribution for the current in an untuned aerial, I have taken it for granted that the relative value of the current at two points in an aerial would be independent of the frequency of the e.m.f., provided the aerial was oscillating as a whole and not in harmonics. The introduction of positive reactance into the aerial circuit is a possible way of balancing an Adcock system. I rather fancy, however, that a resistance balance would in this case be also necessary. I have no proof that balancing the impedances in the manner described in the paper causes symmetry of current distribution about the field coil, but I think it is a very reasonable hypothesis. It is difficult to see about what other point the distribution could be symmetrical than the point which divides the aerial into equal impedances. The type of error to which the system in Fig. 11 is subject is set out in paragraph (3) on page 179. I cannot see that any arrangement of the leads in Fig. 11 could prevent the horizontal length as a whole acting as an antenna which is effectively coupled to the vertical members. The superiority of Figs. 12 and 13 is that this antenna is *broken into short lengths* which are individually too small to have any effect. I do not think I am guilty of using the same field twice over to derive the current in one circuit, as Dr. Smith-Rose suggests. The fact is, the circuits are not the same. Consider the aerial system in Fig. B, which is the combination of a loop and a vertical aerial, and is in many respects analogous to the arrangement questioned by Dr. Smith-Rose, as shown by Fig. C. The current in the limb CE is given by the sum of two currents,

$$I = \frac{Eh}{Z_A} + \frac{2\pi Ehd}{\lambda Z_L}$$

where Z_A = impedance of aerial circuit
and Z_L = impedance of loop circuit,

which is the same as

$$I = \frac{Eh}{Z_A} + \frac{\omega AH}{Z_L}$$

where A = area of the loop. Here we are taking both E and H into account without any redundancy.

Mr. Herd asks how the balanced Adcock compares with the elevated hut Adcock as regards immunity from night errors. No tests of the latter were made with controlled downcoming waves, and as the night tests were comparatively few in number the basis for comparison is scanty. I am inclined to think, however, that the elevated hut Adcock had a slightly better performance than the balanced system. Its characteristic curves would probably be similar to those obtained with the short-wave rotating Adcock of Figs. 19 and 20.

I agree with Mr. Horton that the sensitivity of the Adcock receiver is not as great as that of a closed-loop system of similar dimensions. This is particularly true on long wavelengths. For a system such as that shown in Fig. 10 the ratio of the current in the goniometer field coil to that in a closed loop is given by:—

$$\frac{I_{\text{Adcock}}}{I_{\text{Loop}}} = \frac{hs}{A} \cdot \frac{Z_{\text{Loop}}}{Z_{\text{Adcock}}}$$

Mr. Horton also refers to the screened "U" Adcock as being possibly superior to the balanced Adcock, as explained above. I do not think this is the case. Mr. Horton suggests that a careful comparison of the two systems should be made. I am at present carrying this out and hope to extend the work to a comparative study of the many other possible modifications of the Adcock system. I sympathize with Mr. Horton in the difficulties which he meets in attempting to install an Adcock direction-finder on a battleship.

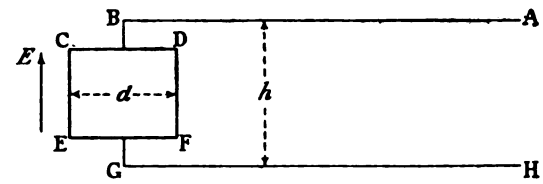

FIG. B.

Lieut.-Col. Lefroy's method of preventing pick-up on the horizontal leads by rotating them is sound in principle, but, on calculation, the amount of pick-up caused by the spacing of the wires is so small that rotation is in the nature of a refinement to which it would not be necessary to resort until the more important sources of error referred to in the paper have been eliminated.

In reply to Air-Commodore Warrington-Morris, the spacing of the vertical aerials certainly affects the pick-up value but not, I think, the accuracy during night effect, since the action of both vertical and horizontal fields is equally affected by any change in this dimension. The most important factor to be considered is the ratio of aerial height to wavelength, which should be made as great as possible. With the central mast arrangement an increase of spacing means a decrease in height; hence the spacing was kept as small as possible. As I have already pointed out in my reply to Mr. Smith, it is not necessary to re-balance on changing wavelength. I am interested to hear particulars of the performance of the Marconi Adcock system. As I have already said, I do not believe that horizontal pick-up can be eliminated or greatly reduced by screening. The "standard deviation" is not the same thing as "average error," but, as defined in Table 1, it is the root-mean-square. One reason why this value was used in preference to the other is that it gives a result free from permanent errors due to site, etc., in which we are not interested.

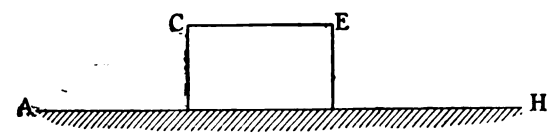

FIG. C.

In hearing from Mr. Eckersley that, in suggesting the

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possibility of balancing the Adcock system, I am forestalled by him by a matter of 13 years, I am rather flattered than discouraged and therefore gladly acknowledge his claim to priority. In developing the completely screened loop direction-finder I have always realized that the same result could be achieved by careful balancing. I think Mr. Eckersley will agree, however, that there is something attractively "fool-proof" about the screened loop which makes it superior for nearly all purposes. Though I should very much like to do so, I have not yet made any tests with the short-wave Adcock over very long ranges. The difficulty is to find time and staff for following up all the avenues opened by a given investigation, and, as Mr. Eckersley says, it will be well to wait until we have more perfect apparatus.

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A WIRELESS BROADCASTING TRANSMITTING STATION FOR DUAL PROGRAMME SERVICE.*

By P. P. ECKERSLEY, Member, and N. ASHBRIDGE, B.Sc.

(*Paper received 4th March, and read before the* WIRELESS SECTION *7th May*, 1930.)

SUMMARY.

The B.B.C. Regional Scheme involves the construction of a number of twin wireless transmitting stations, each capable of radiating simultaneously two programmes on different wavelengths, and each at a power of 50 kW in the aerial. This paper describes the first of these twin transmitters, erected at Brookmans Park, Hertfordshire, to serve London and the Home Counties.

The paper deals with the considerations underlying the choice of sites for such broadcasting stations, the type of power supply adopted, and the factors contributing to reliability in operation, as well as with the wireless transmitters themselves. Detailed considerations underlying the design of such transmitters have been avoided in the anticipation that they will form the subject of a further paper; but, nevertheless, some information in regard to the essential and major points in their design is contained herein.

The paper deals further with the general station lay-out, the building and the auxiliary services, as well as with low-frequency apparatus for supplying modulation to the transmitter and apparatus for testing and monitoring purposes, etc.

Of particular interest is the use of high-tension direct-current machines at this station for the production of high-tension voltage for anode supply to the main power valves. The type of generator developed especially for this work is considered in some detail.

INTRODUCTION.

This paper describes a wireless broadcasting station, installed near London, designed to give two transmissions from virtually the same point on two different wavelengths. Thus a single building contains a power-supply equipment and two identical wireless broadcasting transmitters each arranged to supply modulated high-frequency energy of different frequency to two separate aerial systems.

The station is the first of the " twin " transmitting stations basic to the so-called " Regional Scheme " for British broadcasting. It is true that two different programmes have been radiated during the past two years from the Corporation's station at Daventry, but in this case the two-wavelengths service is given by the use of two separate transmitters in two different buildings a few hundred yards apart; moreover, the difference in carrier-wave frequency is larger.

The implications of the B.B.C.'s Regional Scheme are well known and it suffices to remark here that there will be 5 centres of dual-programme service. The 5 centres of transmission will be located in Scotland, North (Industrial) England, South-East England (based on London), the Midlands, and the South-West (to serve both the West and South Wales areas). Two are completed, Daventry for the Midlands and the station under discussion for London and South-East England.

* Reprinted from *Journal I.E.E.*, 1930, vol. 68, p. 1149.

CHOICE OF SITE.

The site for each of these twin-wavelength transmitters has to be chosen not only in terms of the regions and populations it has to serve, but also in relation to the distribution of the broadcasting service that it supplants.

The selection of the site for the new London station thus had to take into account the fact that, for several years, listeners in London and the neighbouring counties had been accustomed to take their service from a 2-kW station located in the heart of the West End. This location of the old single-wave-service London transmitter had accustomed listeners to a certain distribution of field strength to which they had automatically adapted themselves. Obviously, the measure of dislocation that occurs on changing over from an old to a new station is, to a number of listeners, the ratio of new to old field strength.

A site for the new London station had thus to be chosen on the periphery of the London area. The site, however, could not be so far away from the centre of London that the station erected upon it, albeit of greater power than the old, would give a weaker field strength over all London than the old, and not so near to the centre of London that it would produce intolerable " wipe out " to a great number of persons, owing to the presence of too great a field strength in densely populated areas.

These considerations determined the distance of the new site from the old. It remained to determine the direction of the new site in relation to the old. It would have been quite absurd to have chosen a site along the Thames estuary, eastwards, as much of the service area of the station would have lain over the North Sea; a southerly direction was chiefly debarred by the fact that civil aviation regulations prohibited the erection of masts on the top of the North Downs, which were about the right distance away. A western site would have given a distribution of field strength which would rob Essex and the Eastern counties of good service where strong field strength is required to overcome jamming from the transmissions from sea-going vessels, and, furthermore, would have produced quite unnecessarily strong service in an area already covered by Daventry (5XX and 5GB) transmissions. A northern location seemed, therefore, in all the circumstances to be the best. The northern site was furthermore recommended inasmuch as the existing telephone cables, essential for connecting the studio headquarters in London to the transmitter, were found to be most suitable in a northerly direction.

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Description of Site.

The site finally chosen lies on the east side of the Great North Road a few miles on the London side of Hatfield. It is some 15 miles from the centre of London. Its name is derived from a large estate in the neighbourhood, Brookmans Park. The field which contains the properties, as they have a bearing upon the aerial systems. Thus a 1-kW experimental mobile transmitter was erected on the selected site, in order to prove that the effective height of an aerial erected temporarily between 110-ft. masts was equal to that given by the theoretical assumption that the ground is perfectly conducting.* Certain sites give high earth losses which

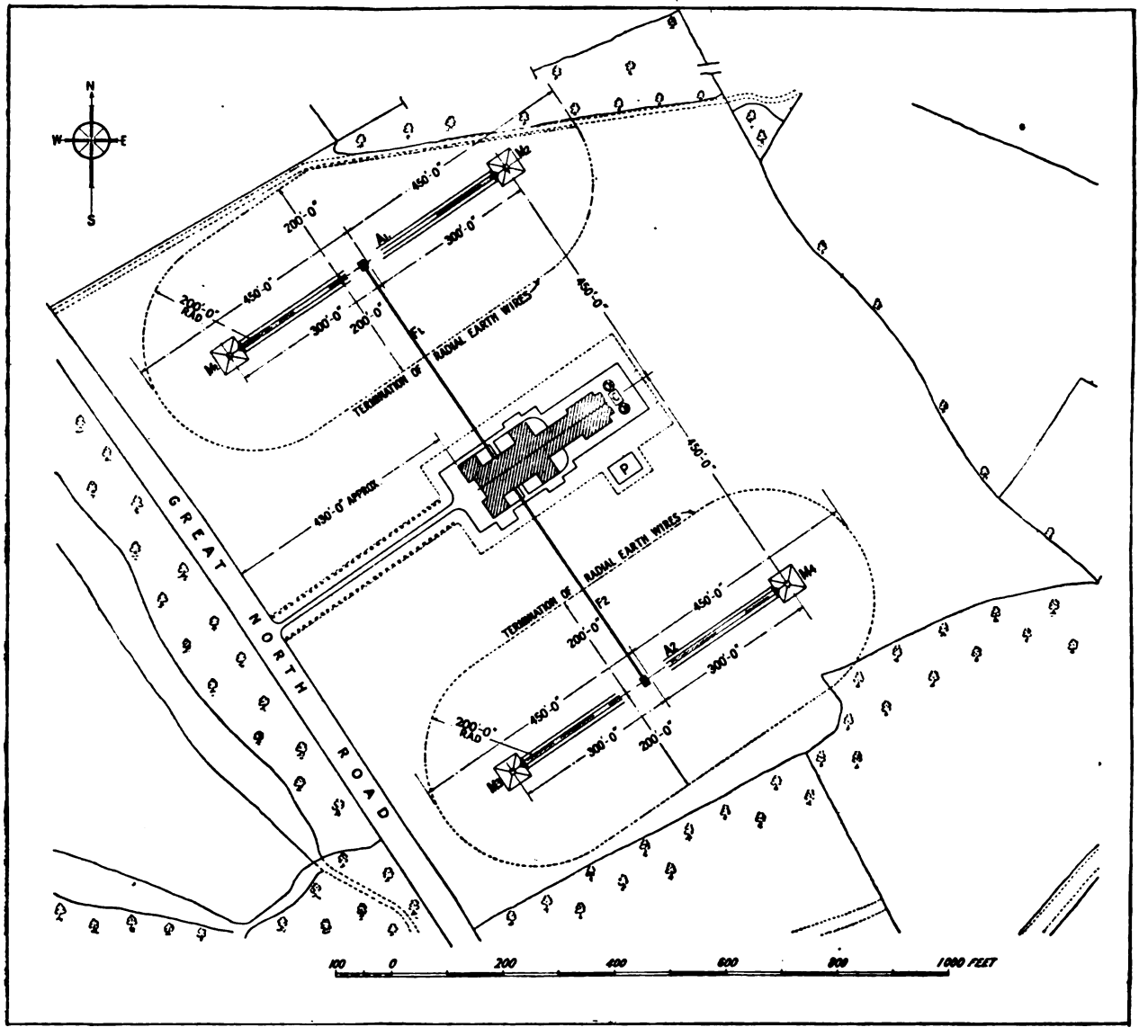

Fig. 1.—General lay-out of station. Block plan of site.

station is 34 acres in extent and is, to all intents and purposes, perfectly flat. Flatness of site is of great advantage in laying out the aerial systems. It also allows the station to be built on one level and so produces a pleasing appearance, in addition to effecting considerable economy.

Telephone cables run close to the site; water, which is required at the maximum rate of 10 000 gallons a day, is available and is taken from the main supply.

Before purchase the site was tested for its electrical reduce the radiation efficiency of aerials erected upon them; it appeared from the experiments made that the site chosen behaved normally in this respect.

The experimental transmitter served the further useful purpose of giving a forecast of the expected field strength in and around the London area. Thus a motor-van containing field-strength measuring-apparatus travelled some 10 000 miles round the site and plotted

* P. P. Eckersley, T. L. Eckersley and H. L. Kirke: "The Design of Transmitting Aerials for Broadcasting Stations," *Journal I.E.E.*, 1929, vol. 67, p. 507; also *Proceedings of the Wireless Section*, 1929, vol. 4, p. 51.

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the field-strength contours given by the temporarily erected mobile transmitter.

STATION LAY-OUT.

Power Supply.

It was thought advisable to install a self-contained generating plant and not to use the public supply mains. It is interesting to note that " shut-downs " of considerable duration at broadcasting stations have frequently been due to failure of the public supply mains. Moreover, it is possible to generate locally at approximately the same cost as, or in some cases a considerably lower cost than, that at which power can be bought. It is obvious that a programme may cost a great deal of money and time in preparation, and it may be of national importance. Failure to radiate the programme, due to lack of power or any other transmitter failure, is therefore a serious matter and it is particularly necessary to eliminate as far as possible all links which may involve long breakdowns. Further, it is desirable to be able to avoid fluctuations in supply voltage. Accordingly, for these two reasons Diesel-engine power units were chosen —generating direct current—so that a storage battery could be used. The battery provides a further factor of reliability in addition to compensating for variations in supply voltage.

General Plan.

The following is a general description of the station; the plant is described in greater detail in later sections.

The station building is so planned that there is a progressive evolution from the crude oil which feeds the Diesel engines (for the power supply) to the transmitters which supply the aerials with modulated high-frequency alternating currents. Thus (see Figs. 1 and 2) the oil tanks (O1 and O2) are located at the rear end of the building. These are sufficient to hold 3 months' supply of oil for the running of the station on full power on both transmitters. Between the oil tanks is the cooling tower (C). The power house contains four Diesel engines, each driving a d.c. generator. The output of the generators is floated across a 2 000-ampere-hour battery, contained in the battery room (B). The output of the power house is taken to switchboards W1 and W2 in the motor-generator room (M). In order to keep the noisy and vibration-making generation plant as far as possible away from the delicate machinery, and yet to maintain a covered communication way throughout the building for the use of the personnel, the battery room (B), packing space (Q), workshop (H) and store (K) are all interposed between the generating station and the motor-generator room. The motor-generator room contains machines which convert the 220-volt supply into the necessary forms for energizing the transmitters, which are located in the transmitter room (T). The output of the machines is led to the switchboard (W3), which is located in the transmitter room. This arrangement is adopted to facilitate operation and economize in personnel. The output from the two transmitters is passed through the two feeder lines (F1, F2) to the aerials (A1, A2). Each transmitter is capable of putting 50 kW into its associated aerial. The feeble low-frequency currents of modulation come from the distant studio in London via a cable line; this terminates in the control room (R2). The currents of modulation are there increased to a sufficient value to be injected into the modulating systems of each of the transmitters.

The valve-cooling arrangements comprise a grid of pipes (P), outside the building, through which the valve-cooling water passes. These stand in a concrete tank, and external cooling water from this tank is allowed to trickle over the outside of the pipes.

Building.

The building itself, which was erected by the Anglo-Scottish Construction Co., Ltd., under the supervision of the B.B.C.'s civil engineer, is, as far as the portions which contain the main technical equipment are concerned, of single-story construction. Many designs were considered in which galleries above the ground-level were to be erected in order to achieve an economy in building space. Calculations showed, however, that the total economic gain was fractional. It is desirable, furthermore, to ensure, as far as possible, freedom from extraneous noise in the transmitter room and so the separation of the machines and transmitters is further desirable. Silence in the transmitter room is of importance not only for the comfort of those operating the equipment, but also because the location of faults due to arcing and sparking of components of the apparatus is much facilitated if there is no extraneous noise. It might, however, have been possible to overcome noise interference by erecting the transmitter in an upper story. Here, again, there are disadvantages, inasmuch as the earthing of the transmitters would have to be made by relatively long earth leads, and it was feared that possible inductive effects between the two transmitters might be incurable. There is of necessity a large amount of ground available, and considerations of actual space are not therefore of paramount importance.

A departure from single-story construction is made in two instances, one where a vault (Y) below ground-level is arranged to contain certain subsidiary apparatus, and the other where the office block has a " first "-floor story. The object of this vault is two-fold. Firstly, it gives easy wiring runs from the machines in the motor-generator room (M) to the output switchboard (W3), and from this switchboard to the transmitters (T1, T2); and secondly it provides space for pumps, tanks, auxiliary batteries, etc. The extremely complicated wiring of the station practically necessitates the vault, because cables run in ducts would be difficult both to inspect and to install.

General Appearance.

An attempt has been made to dispel the illusion that any building connected with engineering is necessarily ugly. Extreme simplicity is the keynote of the architectural finish. Although there is no actual interior decoration in the usual sense of the word, nevertheless an attempt has been made to express in the colour scheme and general arrangement the essentially utilitarian principle of the whole lay-out. The walls of the big rooms are sprayed with white distemper over the common brick. The floors are of red tiles in the engine and

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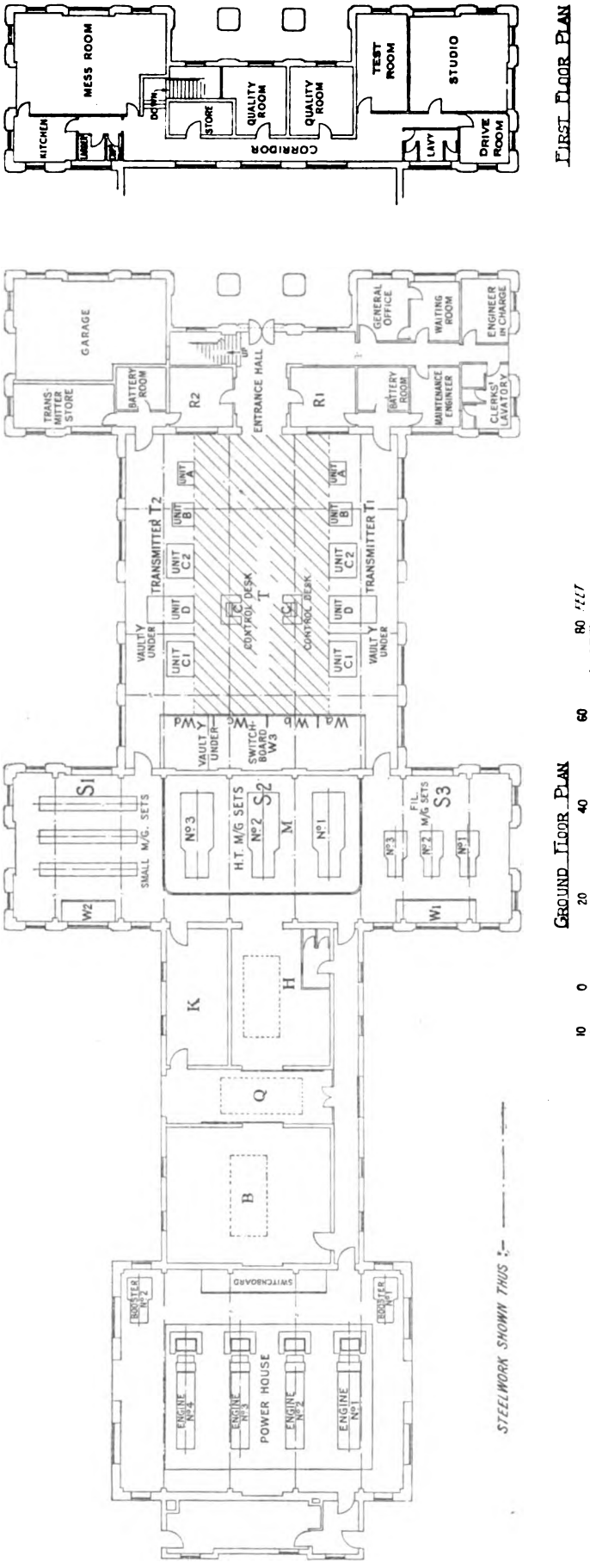

FIG. 2.—General lay-out of building.

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motor-generator rooms, and of red rubber and oak in the transmitter room. Ceilings and office walls are rendered. Ceilings are white but the office-block walls and passages have a grey dado high enough to prevent their showing the dirt inevitably caused by people leaning or brushing against them.

The field on which the building stands is to be left as a grass field. The station, however, is surrounded by a gravel road with a hard core so that heavy transport can deliver loads at the appropriate doors or, in the case of oil supply, to the tanks.

POWER HOUSE.

Each transmitter is designed to give a carrier-wave energy of 50 kW to the aerial, with the transmitter so adjusted as to be capable of 100 per cent peak modulation without distortion. Although actually it was intended to inaugurate the service from this station with a power of about 30 kW to each aerial, the load on the power house, when each transmitter is giving 50 kW to the aerial, has been allowed for and this is approximately 500 kW. It is made up as follows:—

Input to the two main high-tension motor-generators, each absorbing 200 kW	400 kW approx.
Input to two main filament-lighting motor-generator sets, each taking 35 kW	70 kW approx.
Input to grid negative motor-generator sets and motor-generator sets for anode power to drive, separator and modulated amplifier stages on the transmitter ..	9 kW approx.
Motor-driven power-house auxiliaries and motor-driven pumps for valve water-cooler	10 kW approx.
Lighting, cooking, etc.	25 kW approx.
Total	514 kW approx.

Further details of the motor-generator sets mentioned above are given in a later section of the paper.

In order to generate the required power of approximately 500 kW it was decided to use three units, each capable of producing 200 kW. The advantage of using Diesel engines in preference to steam engines for generating power of this description under the conditions which obtain in practice are so obvious as to need no explanation. Actually, four Diesel engines were installed, coupled to d.c. shunt-wound generators supplied by the English Electric Co. and having a voltage range from 220 to 260, the working station voltage being 230. It is intended that only three engines shall run at the same time, the fourth being held in reserve. Fig. 2 shows the general lay-out of the power house.

The size of the power unit was carefully chosen because it allows two engines to be run at just under full load when each transmitter is delivering 30 kW to the aerial, at the same time allowing a margin of about 100 kW with three engines when the whole station is running at maximum load. The decision to use direct current in preference to alternating current was taken principally in order to make it possible to float a storage battery across the mains. This matter, however, is referred to more fully in connection with the storage battery itself.

When considering the type of crude-oil engine to be adopted, the question of vibration was looked upon as being of very considerable importance; first, from the point of view of its possible effect on the transmitters themselves; and, secondly, in view of the possibility of complaints from the owners of surrounding property, since the district is rapidly being developed for residential purposes. For this and other reasons a vertical 6-cylinder engine was eventually chosen capable of developing 50 h.p. per line at 300 r.p.m., i.e. a total brake horse-power of 300. These engines were built by Ruston and Hornsby, Ltd., and are of recent design. They employ so-called " solid injection " of the fuel, and are of the full Diesel trunk-piston type (" full Diesel" is intended here to indicate that no means of igniting the fuel is employed other than the heat generated by compression). Forced lubrication is fitted to all important points, with stream-line filters, integral with the engines, for the lubricating oil.

It is not proposed to give further details of the design of these engines, since they are a standard production intended for general industrial purposes, but the figures under actual working conditions for one commercial grade of fuel may be of interest.

Specific gravity of oil ..	0·915.
Calorific value of oil.. ..	19 200 B.Th.U. per lb.
Full-load consumption ..	0·422 lb. per b.h.p.-hour.
Half-load consumption ..	0·47 lb. per b.h.p.-hour.

Although the type of engine chosen creates comparatively little vibration, nevertheless it was considered necessary to take every precaution to ensure that no detectable vibration should reach the transmitters. With this object in view the engines were mounted on a special bed, constructed in the following way. The four engine units are mounted on a single block of concrete measuring 45 ft. 6 in. × 32 ft. × 6 ft. deep. This block weighs some 600 tons, and the concrete mixture was so specified that the weight should not be less than a fixed amount. The block rests on a raft, but is insulated from it by blocks of material known by the trade name of " Corosil." This material consists of cork mounted in iron frames and given a special treatment to ensure durability. Each frame is approximately 2 ft. 6 in. square, and the cork material is $2\frac{1}{2}$ in. thick. The pressure on the cork is approximately 7·5 lb. per sq. in. Between the foundation block and the retaining walls of the raft there is an air space 3 in. wide which is closed by removable teak boards, fitted flush with the floor of the power house. The special foundations were designed by Christie and Grey, Ltd. Fig. 3 shows a section through the foundation block and raft. This arrangement has proved entirely effective in preventing any detectable vibration from reaching the transmitter; in fact there is no perceptible vibration even in the battery room, which adjoins the power house. The power-house accessories are on standard lines, and for this reason only a very brief description will be given.

Two external tanks, each capable of holding 75 tons, have been provided for the storage of fuel oil, the contents

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of the two tanks together being sufficient to run the whole plant at full load for three months, assuming the normal hours of working. A fuel-oil filter of the centrifugal type is installed for use in the event of the oil being found to contain water or solid impurities. The circulating water for the engines is cooled by means of an external cooling tower of standard design, and the make-up water is taken, so far as is possible, from a concrete pond which receives rain water from the roof of the whole building, the object being to supply soft water to prevent excessive deposit in the water jackets.

The exhaust system was carefully designed to give complete silence 100 yards from the building, and includes a waste-heat boiler capable of supplying sufficient heat by hot-water radiators to all rooms requiring artificial heating.

Battery and Switchboard.

It has already been mentioned that a storage battery is floated across the main supply. This battery, which was supplied by the D.P. Battery Co., Ltd., has a capacity of 2 000 ampere-hours at the 10-hour rate of discharge, and its uses are as follows:—

(*a*) To take the load of one generating set in the event of the sudden failure of an engine or generator. To cover this requirement the battery is rated to give a discharge of 1 000 amperes for 1 hour when discharging to a voltage of 1·75 volts per cell.

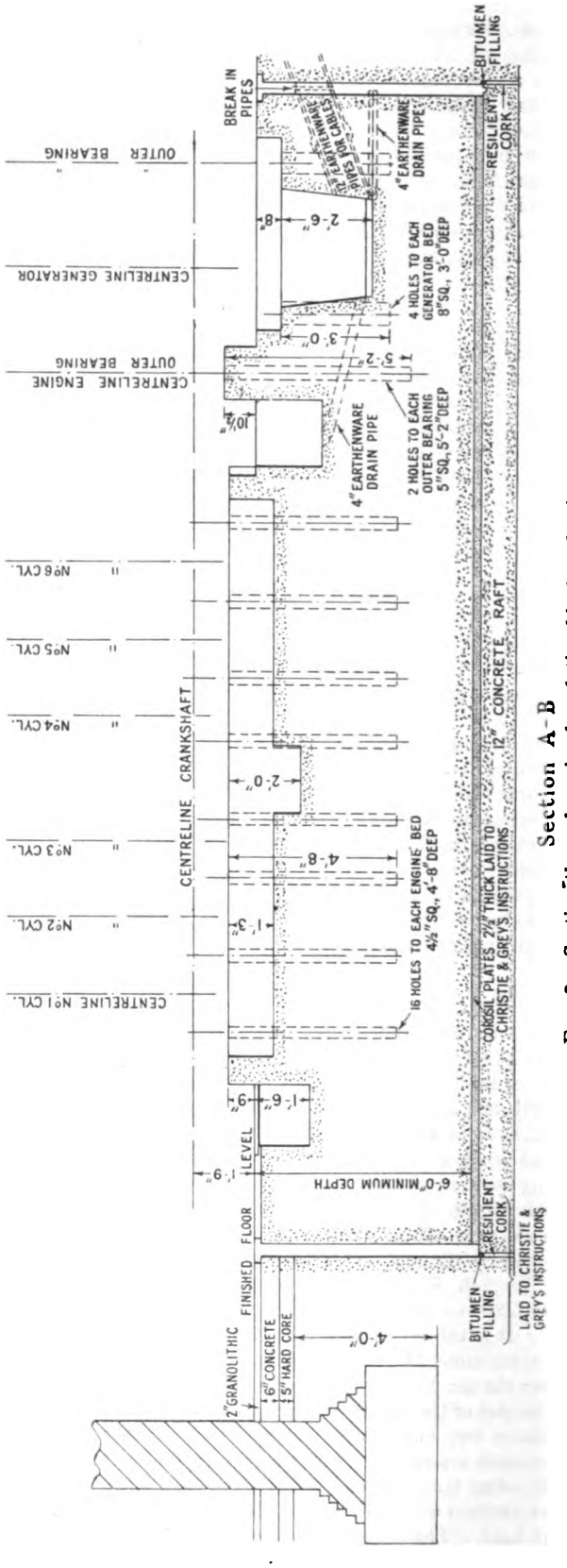
(*b*) To supply power for lighting, etc., and to run auxiliaries and workshop motors when no programme is being transmitted and the load is, therefore, comparatively very light.

(*c*) To assist in maintaining a constant voltage, especially when, for any reason, the load is rapidly varying. In general such occasions will be infrequent, but they arise when one transmitter is being run up while the other is taking the programme, or when some temporary trouble develops on a transmitter or machine.

Requirement (*b*) is sufficient to keep the battery in healthy condition without having recourse to trickle-charging or other special arrangement of this kind.

Two plain shunt-wound battery boosters (one of which acts as a reserve) are provided for battery charging and discharging, no end-cell connections being used. These boosters are hand-controlled by reversing potentiometer-type field regulators. It was not considered justifiable, from the point of view of expense, to install automatic reversible boosters, owing to the fact that the load is practically constant in normal circumstances. In the event of a sudden engine failure, which should, of course, be an infrequent occurrence, the attendant can readjust the voltage in a few seconds when the battery is called upon to take a large load, and the very temporary general drop in voltage will scarcely be detected by a listener.

The power-house switchboard, which was supplied by the English Electric Co., Ltd., is on standard lines and contains the following panels: 4 main generator panels; 2 booster motor panels; 1 battery panel; 2 main feeder



Section A-B

FIG. 3.—Section through engine foundation block and raft.

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panels; 1 power auxiliary panel; 1 lighting panel; and 1 spare panel.

The main feeders running to the motor-generator room are so arranged that the supply to each transmitter can be kept separate and furnished, if necessary, by different generating sets. Thus if experimental work, such as valve-testing, is being carried out on one transmitter while the other transmitter is radiating a programme, one engine set can be isolated so as to supply the load for the test, obviating any risk of affecting the radiated programme by a short-circuit or other accident. This scheme of connections is also particularly useful in the event of some complicated breakdown of one of the transmitters, the cause of which is difficult to trace and which necessitates the frequent running-up of the set. Normally, however, the two sets of busbars and feeders are paralleled. The feeders themselves consist of bare 4 in. × 0·5 in. copper strips run on porcelain insulators along the roof of the passage leading to the motor-generator room.

Motor-Generator Room.

Each transmitter when working on full load requires the following inputs:—

Supply	Volts	Amperes	kW
Main supply to anodes of all water-cooled valves ..	10 000	16	160
Filament-heating for all valves, water-cooled and air-cooled, except the master oscillator	22	1 100	24·2
Grid negative voltage for final power stage	200	4	0·8
Grid negative voltage for penultimate stage	2 000	0·2	0·4
Grid negative voltage for modulator, modulated amplifier and separator	500	0·5	0·25
Anode supply for the master oscillator	2 000	0·06	0·12
Anode supply for the modulator, modulated amplifier and separator	3 500	0·35	1·2

The above are the *actual* inputs for 50-kW working.

Fig. 2 shows the general lay-out of the motor-generator room. It contains three general allocations of space. S1 contains all the small-power machines for grid negative and small high-tension supply, etc. These are very small machines of the order of 1 kW each. They are mounted on pedestals 3 ft. from ground-level and surrounded by a stainless-steel guard which can be quickly removed. This arrangement facilitates the maintenance and repair work, in that the maintenance personnel need not undertake the necessary work in cramped positions. Space S2 is allocated to the high-tension d.c. generators, which take in power at 220 volts and deliver it to the transmitters at 10 000 to 12 000 volts. The remaining space S3 is taken up by the filament-lighting machines, which supply a very large current at a low voltage. The reliability of the machine-room supply is guaranteed in terms of three machines of every kind for two transmitters. Thus there is always a condition where two machines of each kind are working, leaving one spare to be put into service in case of breakdown.

Main High-Tension Supply.

The switchboards controlling the motor side of the convertors are located at W1 and W2. W1 operates the high-tension machines and the filament machines, and W2 the small auxiliaries in the space S1.

The question of the best method of obtaining the main high-tension supply was discussed in a previous paper by one of the authors.* For reasons given in that paper it was decided to use motor-generators in preference to thermionic rectifiers, adopting as simple a design of motor-generator as possible, consistent with reliability. The machines adopted were developed to suit the requirements of this station by Mr. J. E. Calverley, of the English Electric Co. Their maximum output is 160 kW at 750 r.p.m., and the voltage range is 7 000 to 12 000, the machines being capable of giving the full 160 kW at any voltage between these limits. The lower limit of voltage was chosen to cover the possibility of the machines being obliged to work at a comparatively low voltage, on account of so-called " rocky point " effects in the water-cooled valves forming the final power stage of the transmitters. This trouble was somewhat persistent at the Daventry station (5GB) when the design of these machines was under consideration, and at that time the solution was not clear. It was intended to run these machines at approximately 10 000 volts, apart from this consideration, but at the same time it was considered that it might be desirable, at some later stage, to raise the working voltage to 12 000, since the higher voltage gives improved conditions on the transmitter from the point of view of efficiency and linearity of modulated output.

The driving motor is compound-wound, and the generators are separately excited from the station busbars, the fields of the two generators per set being permanently connected in series and controlled by one regulator.

The high-tension voltage is generated in two armatures, each having two separate windings in the same slots and two commutators. All four commutators are permanently connected in series. The voltage per commutator can be varied between 1 750 and 3 000 volts with the aid of the field rheostat so as to give the total high-tension voltage range required.

* P. P. Eckersley: " The Design and Distribution of Wireless Broadcasting Stations for a National Service," *Journal I.E.E.*, 1928, vol. 66, p. 501; also *Proceedings of the Wireless Section*, 1928, vol. 3, p. 108.

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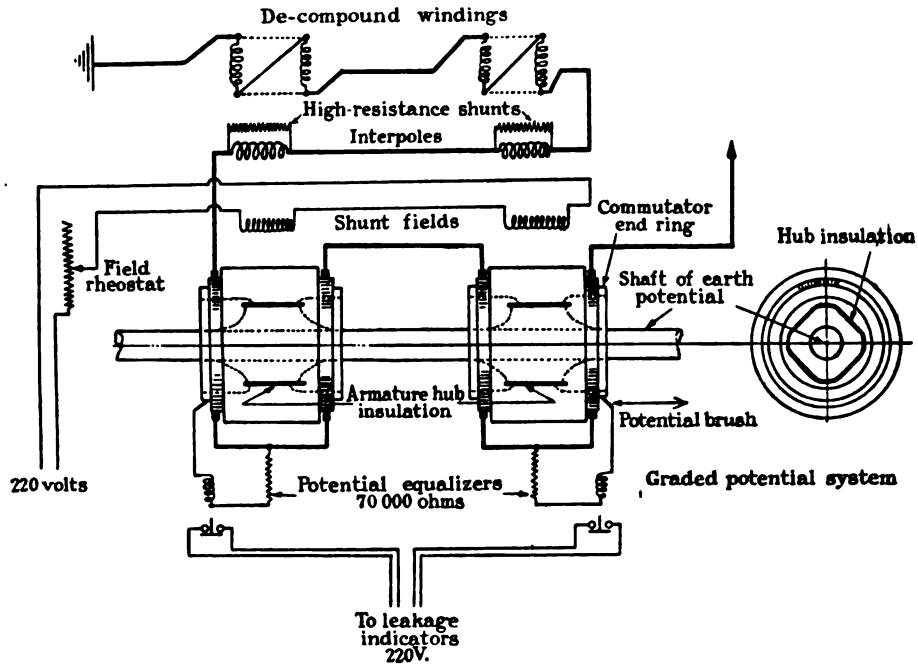

FIG. 4.—Insulation and distribution of dielectric stress in high-tension machines.

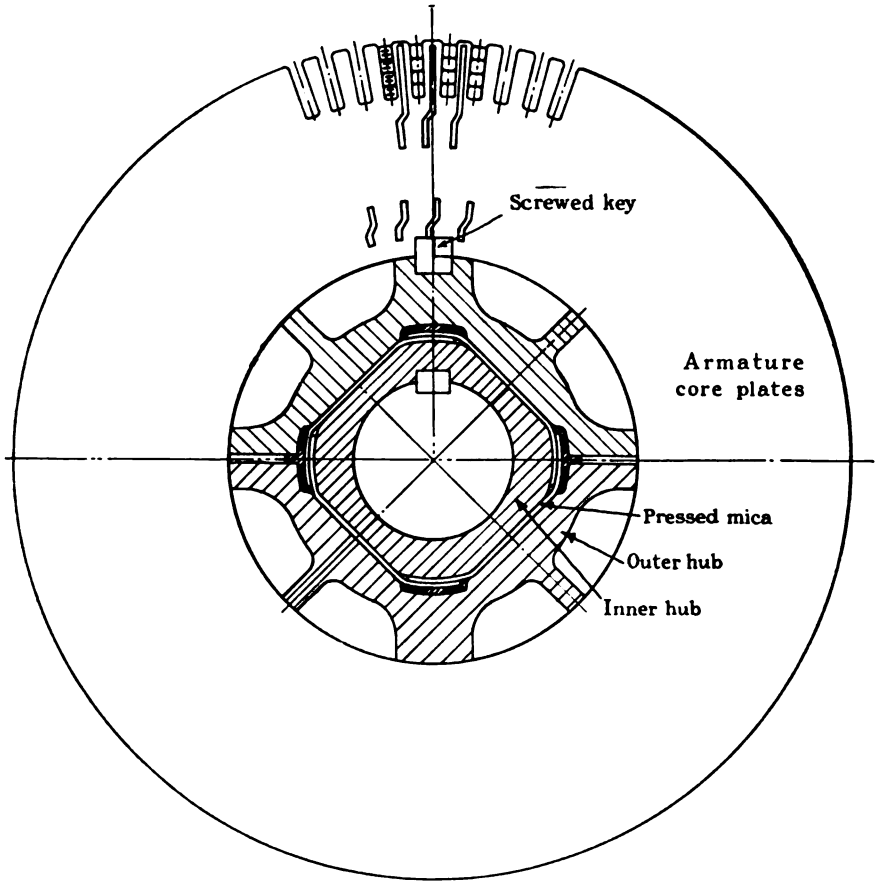

FIG. 5.— Insulation and distribution of dielectric stress in high-tension machines.

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Each machine has four main poles and four commutation poles. The yokes and pedestals are split on the horizontal centre-line to facilitate erection.

The heaviest parts to be handled are the two high-tension armatures and shafts, each of which weighs 2·1 tons.

The shafts of all three armatures (one motor and two generators) in the set are rigidly coupled together and carried in four sleeve-bearings with oil-ring lubrication. Extra precautions were taken in the design of these bearings and pedestals to prevent creepage of oil along the shaft into the machine or on to the commutators. The armatures are mounted on very stiff shafts and were dynamically balanced to ensure smooth running and freedom from vibration.

The insulation scheme and distribution of the electric stress is probably the most interesting feature in the machines (see Fig. 4). Neither the couplings nor the foundations are insulated, the frames, shafts, bearings and bedplate being at earth potential. The English Electric Co.'s graded-potential system was adopted, the object of which is to remove the heavy electric stress from the slot insulation and apply this to a secondary layer of insulation built into the armature hub. The slot insulation consists of pressed mica 0·06 in. thick, and this is subject to a stress having a peak value of 3 000 volts twice per revolution, equivalent to an a.c. voltage of 2 120 at 24 cycles per sec. The stress on the slot insulation is identical in both armatures, though one is operating in series with the other, the system being earthed at the negative side.

The major insulation to earth is incorporated in the armature hub, and consists of pressed mica 0·125 in. thick. This mica is held between solid machined surfaces and projects beyond the armature hub to give a long creepage surface to earth. This layer of insulating material is made up of four overlapping sections moulded to shape and fitted round the inner hub, which is of square section with rounded corners (see Fig. 5). The armature hub in halves was assembled in position on the mica and secured by a shrink-ring at each end. To eliminate the possibility of films of air in any spaces between the mica and the steel, the moulded sections were given a good coating of insulating compound, and this was maintained in a fluid condition until the shrink-rings were in position and the hub complete. The insulating material in this position is very solid and is remote from any appreciable temperature variations. Each hub was tested to the shaft at 25 000 volts, 50 cycles, for one hour before the armature was wound.

The armature core with its hub carries also the commutator hubs, one at each end, so that the whole of the armature and commutators are well insulated from the shaft. The potential of these parts is maintained at its correct value relative to the armature windings by means of a potential equalizer. Two potential equalizers are provided for each pair of generators, one for each armature. They are totally enclosed and are mounted on the bedplate. The connections to the insulated armatures are made through a carbon brush running on an extension of one commutator and ring on each armature core. The potential equalizer maintains the potential of the armature core at a mid-value between the extreme potentials of the two commutators, and definitely determines the electric stress across the slot insulating material, the remaining voltage to earth being applied to the armature-hub insulating material.

The resistance in the potential-equalizer circuit is 70 000 ohms, which limits the fault current in the event of a breakdown of insulating material. If any current exists in this resistance owing to leakage or insulation breakdown, a relay connected in series will close two contacts on a low-potential circuit and operate the leakage-indicator gear.

All highly-inductive windings which are connected in the high-tension circuit, such as interpole coils and series windings, have reinforced insulating material on the end turns of each coil to safeguard them against breakdowns between turns due to surges or impulses. The interpole windings are provided with non-inductive shunting circuits to bypass any audio-frequency currents which may exist in the system.

The interpole coils and series coils are connected at the earth side of the armature, but are insulated for a high-

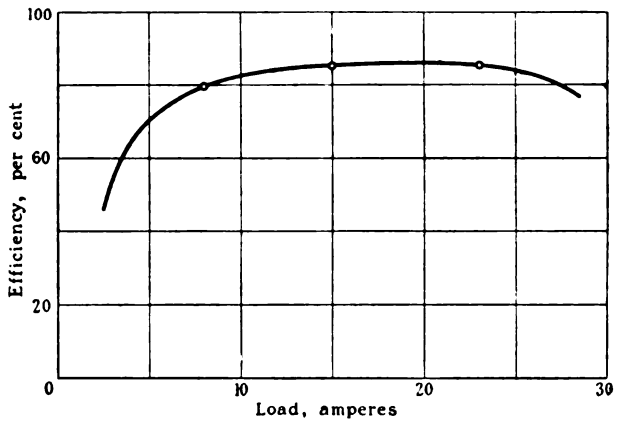

FIG. 6.—Overall efficiency of high-tension sets.

pressure test of 5 000 volts (a.c.) owing to the fact that in certain circumstances higher voltages to earth than normal may exist.

Insulation Tests.—The following high-pressure tests were carried out on the finished machines, using alternating potential from a 50-cycle circuit. All these tests were of 1-minute duration.

Armature windings to armature core ..	7 500 volts
Armature core to shaft	20 000 volts
Series and interpole windings to frame..	5 000 volts
Series winding to shunt winding ..	5 000 volts
Shunt winding to frame	2 500 volts
Potential equalizer and relay coil to bedplate	20 000 volts

Considering that the slot insulating material and the armature-hub insulating material are electrically in series, the total final test pressure between the armature windings and earth is equal to 27 500 volts (a.c.).

Efficiency and Temperature-rise.—The overall efficiency of these high-tension sets is shown in Fig. 6, from which it will be seen that the efficiency over the working range is from 85 to 86 per cent.

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The machines have been designed to comply with B.S.S. No. 168—1923, as regards temperature-rise. The figures specified therein are not exceeded even if the

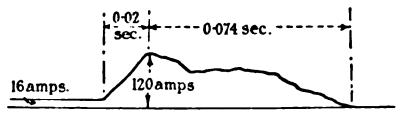

FIG. 7.—Oscillogram showing current when generator is short-circuited (with circuit breaker).

maximum rated voltage of 12 000 and the maximum rated current of 23 amperes are taken together.

Voltage Characteristics.—Owing to the possibility of

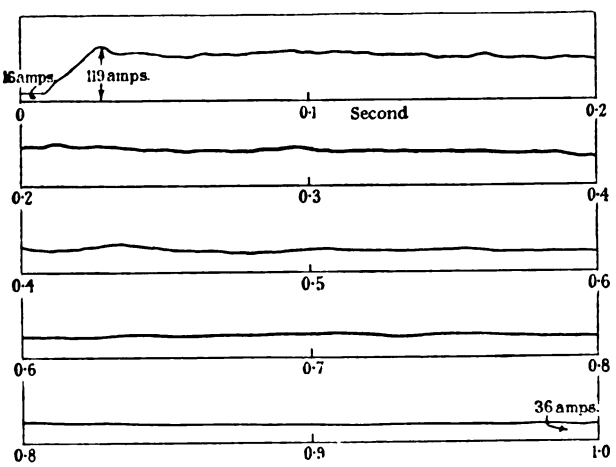

FIG. 8.—Oscillogram showing current when generator is short-circuited (without circuit breaker).

"rocky point" effects in the main transmitting valves, it was considered desirable to give the machines some form of inherent current-limiting characteristic, and for this reason the generators are heavily de-compounded with their own current. It was recognized that de-compounding could not be instantaneous in its action and prevent the first rush of current on a dead short-circuit, but as the smoothing system includes a high-value choke coil (5 henrys) in series with the machine, the rate of increase of current due to a short-circuit external to the choke and machine could be controlled and made sufficiently small to permit the de-compounding to keep pace with the current-rise.

The total inductance of one pair of machines in a high-tension motor-generator set is about 1·1 henrys, and the oscillograms in Figs. 7 and 8 show that without the choke in circuit the initial rate of increase in current on short-circuit at the terminals is in the neighbourhood of 5 000 amperes per sec. The de-compounding winding has little effect at the instant of short-circuit, but after about 0·02 sec. (during which time the current is increased by about 100 amperes) the de-compounding becomes effective (Fig. 7).

With the 5-henry choke in circuit, the initial rate of increase of current is reduced to about 1 500 amperes per sec. and the current will not rise to more than about 35 amperes before the voltage begins to fall, due to the action of the series de-compounding windings. As this is the final short-circuit current, when the circulating current approaches this value it remains practically steady whilst the voltage falls to zero. This scheme has been found to be quite effective in service, and a short-circuit in the transmitter room does not cause any appreciable disturbance on the commutators of the motor-generator set.

On account of the wide range of current for which these machines are designed, the series windings on each pole are divided into two sections, and these can be connected for use on the higher current range so as to make two complete parallel circuits on each generator field. A further reason for this splitting of the series

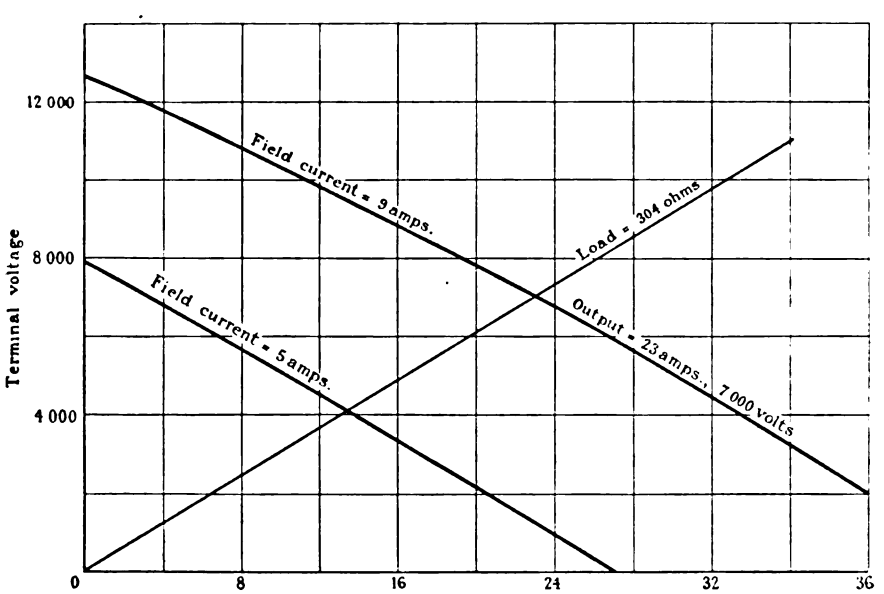

FIG. 9.—Load characteristics of high-tension machine with half de-compounding.

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windings and putting half on each side of the shunt winding is to keep down to a low value the magnetic leakage between the shunt and series windings and to render the de-compounding windings quicker in their action.

always build up on load, and in the event of the circuit breaker operating it opens the shunt field circuit and eliminates the excitation.

The steady short-circuit current of the machine is about 36 amperes, and the curve shows how the voltage

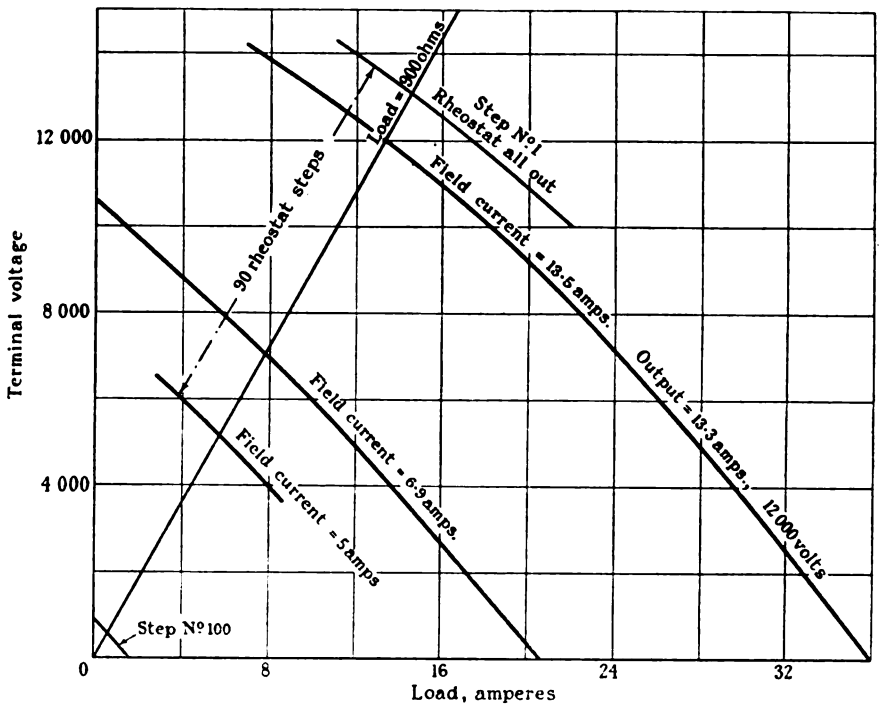

FIG. 10.—Load characteristics of high-tension machine with full de-compounding.

Links, provided on the lower portion of each magnet frame, permit of these sections being connected in series or in parallel to suit the current range required.

Figs. 9 and 10 show the characteristics of the machine with each arrangement of the connections. Fig. 10

falls as the current increases. The load on these machines is practically a dead resistance load for any one adjustment of the transmitter, and under these conditions the field control is very precise and effective. Under the control of the field rheostat the machine will

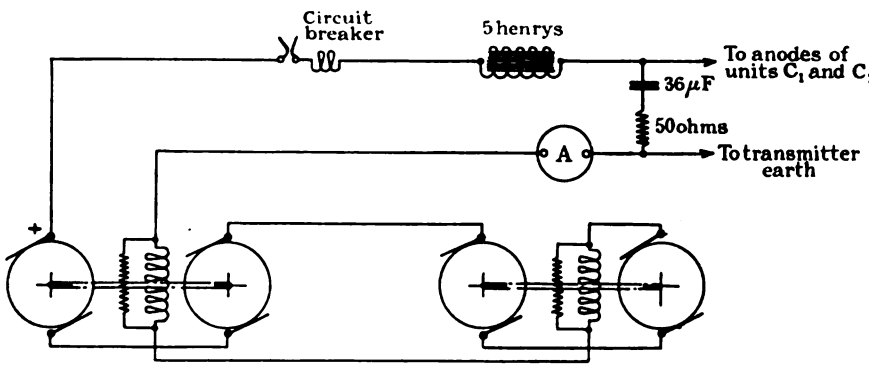

FIG. 11.—Diagram of connections of high-tension machine with smoothing set.

illustrates the characteristics of the set with the de-compounding windings in series. This is a normal working condition for an output on the 9 000- to 12 000-volt range. It will be noted that the no-load voltage-rise is in the neighbourhood of 25 per cent, though this voltage-rise never occurs in practice because the machines

build up its voltage, following the straight line which is shown in the diagram to correspond to the resistance load of 900 ohms, which would absorb 13·3 amperes at 12 000 volts.

De-compounding curves for other values of the field current are also shown in Figs. 9 and 10. These all take

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the same shape and only differ in value with different values of the field current. Under these conditions the machine will build up along the straight line under the control of the field rheostat, and the increments of voltage per step of the rheostat are about 70 to 80 volts.

Short-circuit Characteristics.—Short-circuit tests were taken on these machines with and without the circuit breaker (see Figs. 7 and 8). The short-circuits were taken across the terminals of the machine. Fig. 7 shows the change and duration of the current with the circuit breaker in circuit. This curve was taken when the machine was delivering 16 amperes at 10 000 volts, and the maximum current recorded was about 121 amperes. This current commenced to decrease in 0·02 sec. owing to the de-compounding, and was interrupted by the circuit breaker, which brings the current to zero in about 0·095 sec. The circuit breaker is not of the special rapid type, since the impedance of the systems, together with the effect of the de-compounding, prevent any excessive current; it is, in fact, fitted with an adjustable time-lag device.

Fig. 8 shows a similar test without the circuit breaker in series. The wave is of similar shape, but instead of being interrupted by the circuit breaker the current is gradually brought down to its steady short-circuit value by the action of the de-compounding winding. The peak value of the current in this case is much the same as before, reaching its maximum value at 0·02 sec. from the instant of short-circuit. At this time the de-compounding winding becomes effective and gradually reduces the voltage and circulating current, until at the end of 0·95 sec. the current is steady at its short-circuit value.

Fig. 11 shows a diagram of connections of the machine with its associated smoothing circuit.

An oscillograph test was taken to determine the amount of the commutator ripple. It was specified that this should not exceed 1 per cent of the working voltage on either side of the mean, without the smoothing system in circuit. The ripple as actually measured in this way was approximately 0·08 per cent.

There are three of these machines installed, and each machine is enclosed by a separate iron grille the gates of which cannot be opened without closing down the machine. The main smoothing chokes, one for each machine, have a value of 5 henrys each, are oil-cooled, and are also mounted in these compartments.

Motor-Generators.

For Filament Heating.—These machines, which were also supplied by the English Electric Co., Ltd., are entirely straightforward and the only item in the specification which is in any way unusual is in connection with ripple. The maximum amount allowed was 1 per cent above or below the mean of the working voltage; the figure as measured by an oscillograph was approximately 0·5 per cent. As in the case of all the motor-generator sets for the various supplies to the transmitters, three machines were installed, one for each transmitter and one in reserve. The rating of the filament-heating motor-generator sets is as follows:—

Voltage range	15 to 30
Current	1 300 amperes
Speed	570 r.p.m.

The results of the tests were as follows:—

Efficiency	76 per cent
Highest temperature-rise	30 deg. F. (generator commutator)

For Anode Supplies to Separator, Modulator and Modulated Amplifier.—These machines were supplied by Messrs. Haslam and Newton, Ltd., and were given a specially liberal rating owing to the heavy duty required. The rating is:—

3 500 volts and 0·6 ampere; measured ripple 0·2 per cent.

The same maximum ripple was allowed as for the main h.t. machines, viz. 1 per cent.

For Master Oscillator High Tension.—These are rated at 2 000 volts and 200 mA; measured ripple 0·57 per cent.

For Grid Negative Supplies.—These are rated as follows:—

Main amplifier 250 volts, 4 amperes; measured ripple 0·5 per cent.

First power amplifier 2 000 volts, 200 mA; measured ripple 0·57 per cent.

Modulator and separator 500 volts, 500 mA; measured ripple 0·3 per cent.

TRANSMITTER-ROOM EQUIPMENT

Lay-out.

The transmitter room is 73 ft. 10 in. long, 60 ft. wide and 26 ft. high (see Fig. 2). The two transmitters T1 and T2 face one another, leaving a large space in the middle of the room. The switchboard is at right angles to both transmitters. The previously mentioned vault Y runs so that there is a corridor-shaped space below the transmitters and the switchboard. The height of the vault allows a man to walk upright, and the whole space is sufficient to take all wiring runs from machines to switchboard and from switchboard to transmitters. The main wiring was carried out by Drake and Gorham, Ltd. The vault contains, in addition, tanks, pipe runs, auxiliary batteries and pumps necessary for the valve-cooling plant.

Switchboard.

The switchboard on which the outputs of the machines are centralized and controlled before being passed on to the transmitters is divided into four partitions Wa, Wb, Wc, Wd. Wb and Wd deal with the output of the small auxiliary machines in space S1 (see Fig. 2) for grid negative, power supply for the drive, and so on. Wc contains the section for the output of the high-tension d.c. generators in space S2 (see Fig. 2). Wa deals with the output of the filament-supply machines, which are themselves located in the motor-generator room in space S3 (see Fig. 2).

It is obviously essential to arrange the switchboard in such a way that the output of any of the three machines in each group can be diverted to either of the two transmitters. The methods used are straightforward and do

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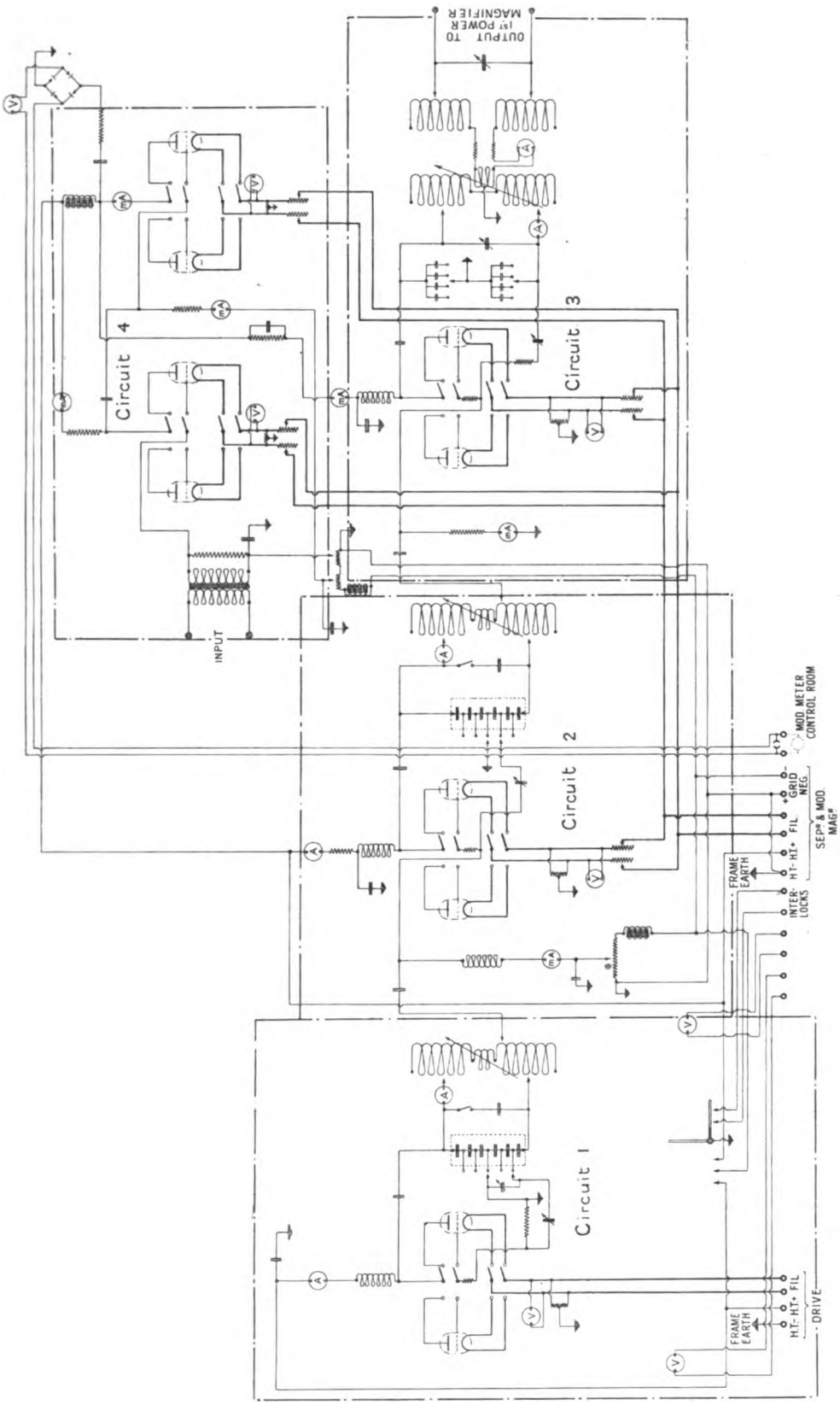

FIG. 12.—Schematic diagram of unit A.

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not merit description in a general paper. The arrangement of the high-tension switchgear which is found in the centre of the switchboard ensures absolute protection of the operating personnel, and none of the live parts comes in contact with the actual face of the switchboard. The instruments are recessed and are protected by a glass window flush with the face of the switchboard. The circuit breakers must of necessity be in such a position that, when they interrupt the current, the ensuing flame does not burn any part of the apparatus. To this end the breakers are placed above the centre part of the switchboard. The front magnetic shields of these breakers are at full positive potential and it is, therefore, necessary to protect them with an earthed expanded-metal screen so as to be absolutely certain that the personnel shall not come into contact with the live parts of the breakers, when carrying long pieces of conducting or semi-conducting material, etc. The smoothing condensers are installed behind the switchboard. The switchboard and doors cut off all access to the components at high tension.

Transmitters.

(*a*) *General Design.*—A previous paper * outlined the reasons for the adoption by the engineering branch of the B.B.C. of what is known as "low power" rather than "high power" modulation as the basic design for the transmitters themselves. Low-power modulation involves generating oscillations of the required high frequency, modulating them at comparatively low power and then magnifying the resultant by as many stages of high-frequency amplification as is necessary.

Each transmitter proper contains four units, as follows:—

The A unit contains all the circuits for producing modulated, relatively low-power oscillations, the fundamental or middle frequency being maintained sensibly constant even under conditions of maximum distortionless modulation.

The B unit increases the power of these modulated oscillations.

The C and D units comprise apparatus which increases the output of the B unit to the value required for radiation.

The A unit contains four separate shielded circuits, the drive, separator, modulated oscillation amplifier and modulator. A schematic diagram of the A unit is shown in Fig. 12. Circuit 1 is a conventional thermionic-valve oscillation generator. The frequency of oscillation of this circuit determines the frequency of the aerial carrier-wave current. To prevent frequency-change the drive circuit must work into a constant load. This condition is achieved by using a so-called "separator," circuit 2. The output of circuit 2 energizes circuit 3, the modulated oscillation generator. This latter acts to magnify the oscillations passed to it from the separator. The modulator (circuit 4) follows the conventional choke or Heising system, and modulates the intensity of these oscillations in sympathy with the low-frequency currents of modulation fed to it from the local control rooms.

The output from unit A is passed to unit B, which contains a push-pull connection of two water-cooled valves designed to magnify the modulated high-frequency currents fed to it from unit A. The output of unit B is passed on to the identical units C1 and C2. The latter contain banks of water-cooled valves, one unit taking one side of a push-pull circuit and the other the other side. Unit D contains a closed circuit consisting of inductance and capacity which is set into full-power oscillation by connection to the valves in units C1 and C2. Arrangements are made for neutrodyning the various circuits. Schematic diagrams of connections are shown in Figs. 13, 14 and 15.

It will be seen that the system comprises essentially a high-frequency amplifier which magnifies the modulated high-frequency currents generated in unit A. The output of the last stage is passed to the aerials via feeder lines.

It is not thought necessary to include a full description of the transmitter here, because otherwise the paper would be too long, and it is anticipated that a further paper will be written, but, the general lines having been indicated, it will perhaps be of interest to consider some salient points in more detail.

(*b*) *Construction.*—The transmitters were designed by the B.B.C. engineering branch, and were constructed by Marconi's Wireless Telegraph Co., Ltd. They are based on data obtained by the design and operation of an experimental station at the Corporation's transmitting centre at Daventry, familiar to listeners as station 5GB. The transmitters are thus the outcome of experience with an existing experimental "lash-up" which has been tested under conditions of service for over a year.

An attempt has been made throughout to design the apparatus so as to ensure absolute safety for the personnel operating the station. Thus each unit contains all its associated apparatus in an earthed aluminium screen, but where it is necessary to read instruments and keep the apparatus under observation, glass doors give this facility. The cases are so made that access to the apparatus is perfectly easy by opening the doors, but no door can be opened without automatically cutting off the power, as the door handle is connected to a switch which cuts off the supply when the handle is turned to allow the door to be opened. The closed metal construction has the further advantage of minimizing stray fields.

Great care has been taken to ensure that the components of the apparatus shall be sufficiently accessible, and, furthermore, to facilitate the quick change of faulty components these are in most cases not screwed down, but rest on spigots. Thus to remove a component it is only necessary to disconnect the various wires which run to it, lift it out and substitute a new one.

The essential circuits each contain a meter for registering the current or voltage essential to its operation. Thus, among other quantities measured, every valve grid and anode has a separate meter in order that the operating personnel may have continuous assurance that all the valves of the team are behaving identically.

Water-Cooling.—In the Daventry high-power long-wave station, designed five years ago, the insulation of the valve-cooling water, which must, of course, come in contact with the anodes, which are at several thousand volts above earth potential, is obtained by running it through a perforated plate somewhat similar to the rose

* P. P. Eckersley: "The Design and Distribution of Wireless Broadcasting Stations for a National Service," *Journal I.E.E.*, 1928, vol. 66, p. 501; also *Proceedings of the Wireless Section*, 1928, vol. 3, p. 168.

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of a watering can. This breaks up the water into fine drops and achieves practically perfect insulation. By allowing the water to run into the valve through a spray and go out from it also through a spray, sufficient insulation is achieved. There are, however, considerable maintenance troubles with such a system, and small particles of grit inevitably get into the water and are inclined to choke up the fine holes in the spraying plate. In the station being described a more conventional method of running the water into the valves through a coil of large rubber pipe, and forming an exit for the water in the same manner, has been adopted. There is, with this system, some direct-current loss due to conduction of current to earth via the water, and there is furthermore some loss due to high-frequency currents feeding into the pipes as a capacity to earth. However, neither trouble is serious. The actual valve-cooling water is fed into the valves by gravity from a roof tank. It passes, with rubber-pipe entrance and exit, over the valve anodes and thence to a tank in the vault, whence it is pumped through the cooler into the roof tank. The water which comes in contact with the valve anode is thus in a totally enclosed system, and this allows water having the correct degree of hardness to be used. The cooler itself consists of a bank of tubes mounted over an open concrete pond, water from which is sprayed over the outside of the pipes by means of a pump situated in the vault. The cooling system is so designed that the maximum temperature of the valve-cooling water leaving the valve jackets shall not exceed 110° F. under summer conditions. A separate bank of tubes is installed for each transmitter, but the external water system is common to both. This portion of the plant was supplied by Messrs. Rosser and Russell, Ltd. Instruments of the remote-indicating type give a continuous reading of rate of flow and temperature. Great care has been taken to ensure very good cooling of the valves, seeing that some trouble with valve flash-over was experienced on the experimental station. This latter point needs some further emphasis.

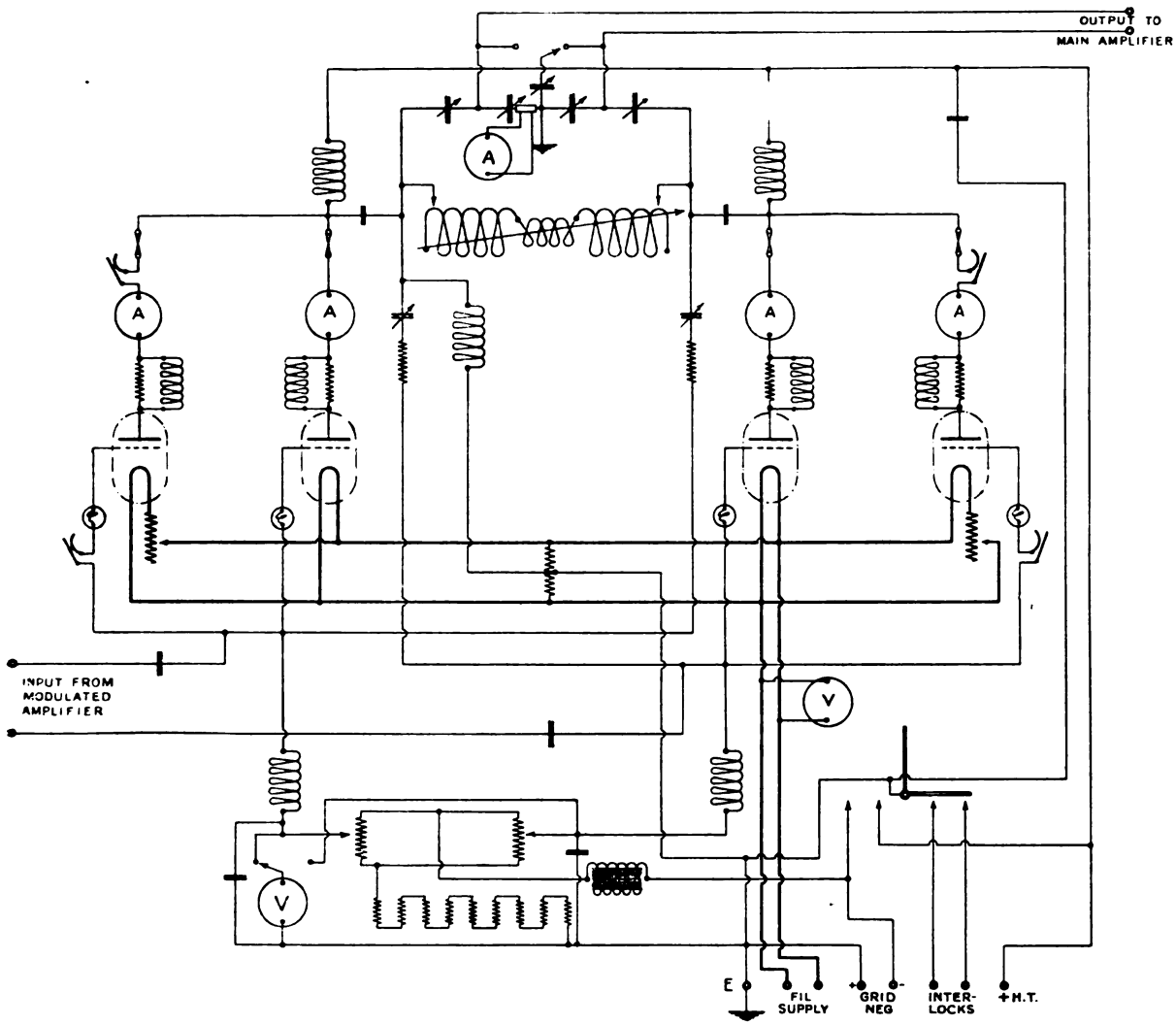

FIG. 13.—Schematic diagram of unit B.

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Valve Flash-Over.—The chief difficulty met with at the experimental station was the frequent occurrence, soon after the set had been passed for service, of valve flash-over or " rocky point " effect. These flash-overs caused complete temporary shut-down of the station when d.c. generators were used, but did not cause such serious interruption when rectified alternating current was used; the latter naturally has a greater effect of limitation of current when valve flash-over occurs. While a full discussion is reserved for a further paper, it is interesting to note that the trouble was eventually cured, but there has been no opportunity to determine whether this was due to improvement in valve manufacture or to a rearrangement of the circuit. Originally the 5GB

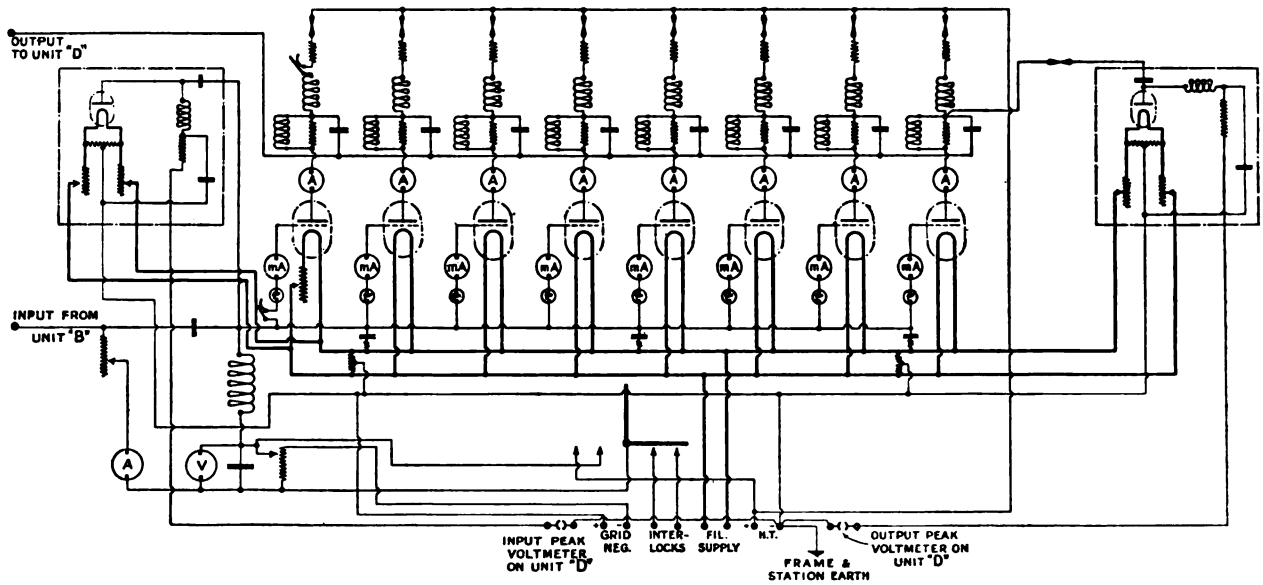

FIG. 14.—Schematic diagram of units C1 and C2.

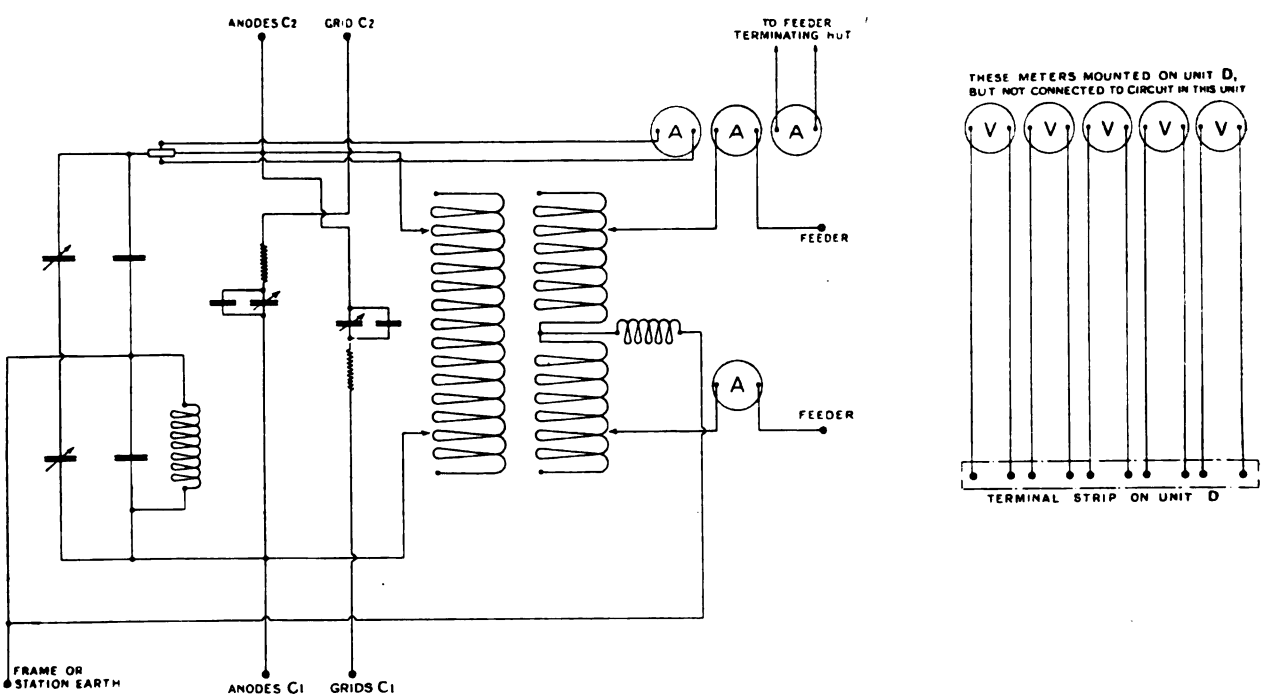

FIG. 15.—Schematic diagram of unit D.

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circuit was conventional and worked on to the closed circuit in the usual "top feed" connection. It was noticed that flash-over occurred only when many valves were connected in parallel, and not on the penultimate stage where only two valves in parallel were used. The circuit was changed, therefore, to one in which each valve was fed individually through a separate air-cored choke and in which each valve anode contained a resistance. The object of individual feed was to try to arrange that each of the many valves energizing the last stage could not be influenced by the other valves. This appeared to have an influence in curing the trouble. The resistance was useful in checking incipient oscillations of much higher-frequency value than that which they were intended to magnify, and in minimizing the short-circuit current when valve flash-over occurred. It was finally decided to use a push-pull connection as well as individual valve-feed. Since this connection has been used no flash-overs have been experienced at the 5GB Daventry station, and the new transmitters at Brookmans Park appear to be perfectly stable. The interesting point to notice, however, is that the new circuits were used with new valves of better design (from the valve makers' point of view) and it is still undecided whether the new circuit or the new valves banished the trouble. There has been no opportunity, since 5GB is now a service station, to test the old circuit with the new valves or, much less, the old circuit with the old valves, since these are now either destroyed or in use on other work. The new transmitters, however, are designed not only with a circuit which is apparently stable, but also with every kind of precaution to minimize shut-down if and when "rocky points" occur (see former section of paper).

Quick Valve Replacement.—Great precautions have been taken to ensure reliability of performance. To this end, where there is one particular valve in a circuit which, if it failed, would put the rest of the apparatus out of action, arrangements have been made for throwing in a spare valve without cutting off the power from the particular unit. If, for instance, either one of the valves of unit B fails, the operator of the station can bring a new valve into operation in perhaps 15 seconds or less. To this end, a switch is mechanically arranged so that when the spare valve filament is brought up to full brightness by turning a handle operating a filament rheostat, a spring switch is released which connects the anode and grid of the spare valve. This operation does not, therefore, involve the necessity of cutting off power and water supply, taking out the faulty valve from its water jacket, substituting a new valve and switching on water and power again, as indeed has to be done with the older designs of equipment. The method of quick valve replacement is applied to all units. A carbon tetrachloride type of fuse in the anode circuit of each water-cooled valve serves to disconnect the valve which has become faulty, if indeed the fault has caused a short-circuit.

Measuring Instruments and Remote Control.—It is obviously necessary to ensure that the correct values of current and voltage are supplied to each component of the set, and continuous watch must be kept to ensure correct adjustment of supply voltages. Each transmitter has thus a remote-control table (see C, Fig. 2), and it is intended that the engineer shall sit at this table and keep a continuous watch on the various measuring instruments, which are located not only on the set but also immediately under his eyes on the table itself. Variations are taken care of by handles on the control table which regulate the voltage of supply by field-excitation regulation of the machines concerned. The quantities under direct control of the operating engineer and controlled by handles on the control table are filament voltage and high-tension supply voltage; other quantities also liable to variation can be controlled from the main output switchboard. Controls are also provided for tuning and adjustments on the units themselves.

The transmitter is theoretically capable of 100 per cent distortionless modulation, but distortion must inevitably occur when the input modulation exceeds this quantity. Voltmeters are connected in such a way that the operating personnel can, at any moment, see the amount of modulation. These meters are calibrated with instruments in the local control rooms, so that the general level of the ingoing modulation current can to some extent be regulated by local personnel. (The final authority of control is, however, given to the controlling engineer in the control room local to the studio, and the transmitting personnel, although they are given facilities for altering the degree of modulation, are under orders to make this general rather than momentary.)

Efficiency of Output Stage.—The main power valves are allowed to be operated with grid current, and precautions must therefore be taken against the occurrence of distortion due to variable loads on the input. Great loss of efficiency would occur if the positive grid region were not used. The effect of grid current is overcome by shunting a resistance from grid to filament of the power valves, of such a value that it is small compared with the effective resistance of grid to filament. In the static case, i.e. before modulation, the efficiency of the last stage is 28 per cent. This efficiency increases and decreases with modulation.

Earthing.—The earthing of the transmitter is to a grid of wires which were buried in the ground when the foundations of the building were being laid. There is thus what is called a "general station earth" which is separate for each transmitter. Great importance attaches to the symmetry of the earthing arrangements of the high-frequency units, and great care has been taken to avoid unequal earth paths. The high-frequency earths of the radiating systems are discussed later in the paper.

Feeder Lines.—The high-frequency feeder lines are connected to a high-frequency transformer circuit, the primary of which is formed by the inductance of the closed circuit in unit D, while the secondary is connected to the feeder lines. The feeder line terminates in the transformer house.

Aerial Systems.—The masts (M1, M2, M3 and M4) were supplied by the Foundation Co., Ltd. As will be seen from the general lay-out plan in Fig. 1, they are placed as far as possible from each other and from the building. The designers of the station were faced with the necessity

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of using relatively short masts. The Air Ministry considered that, in view of the amount of flying that takes place round the area, it would be dangerous to build masts of greater height than 200 ft. With such dimensions the problem is best solved by erecting four masts and placing them, as mentioned before, at the maximum possible distance one from another. Masts M1 and M2 support one aerial, and masts M3 and M4 support the other. A distance of 900 ft. separates one aerial from another. The object of this large separation is to ensure that there shall be no heavy induced currents in one aerial system from the other, and to try to arrange that the induced currents in any mast from the aerial it supports shall be so small as to produce little distortion of the horizontal polar diagram. The masts are insulated from the ground chiefly to give such a flexibility of adjustment that the mast can either be earthed or insulated, depending upon the wavelength used and the amount of distortion of the polar diagram found in practice. It is of interest to note that tests undertaken on specimen porcelain insulators, such as are used on the masts, showed that they were easily capable of withstanding a high-frequency voltage of 10 kV at 1 000 kilocycles per sec. without undue heating. When sprayed either with distilled or " hard " mains water no brushing was apparent. In practice very thin lead washers are fitted between the porcelains and between the mast foot and the end insulators, in order to obtain as intimate and evenly distributed a contact as possible. This was done to avoid local heating, which was observed on test if intimate contact between the metal and the porcelain and between each pair of insulators was not ensured. The technique of distant mast spacing naturally involves the use of a feeder line to conduct the high-frequency energy from the transmitter to the appropriate aerial.

Earth Systems.—A separate high-frequency earth system is arranged symmetrically under each aerial. Each earth consists of a number of copper wires radiating out from each feeder hut—one every 10°—the wires being buried about 9 in. below the surface and of such length that the earth system stretches out to about $\frac{1}{4}$ wavelength beyond the dimensions of the aerial (see Fig. 1). The wires do not terminate in plates.

Office Block.

The office block (Fig. 2) contains the two control rooms, with their associated battery rooms, offices and rooms for subsidiary apparatus. The offices comprise a general office, an office for the engineer-in-charge, and an office for the maintenance engineers. The ground floor is completed by a garage, waiting-room, hall and lavatories. The upper story comprises a staff mess-room to which are attached a larder and kitchen, and two listening rooms where the quality of the transmission can be checked on loud-speakers. A gallery for access to rooms on the upper story overlooks the transmitter room and gets borrowed light therefrom. The upper story also contains a test room, a studio, and a small room to be used for tuning-fork gear. A detailed description of the above follows.

Listening Rooms.

Two rooms have been set aside for routine quality-checking by loud-speakers and they are intended primarily for use by the engineer-in-charge of a shift. It was considered necessary to provide two such rooms to enable simultaneous observations of the quality of each transmitter to be made by two listeners under similar conditions. There are, in fact, numerous occasions on which it is desirable to use both these rooms at the same time, and for this reason it was not considered advisable to provide only one. The equipment of each room consists of a loud-speaker which normally takes its input from the check receiver, but arrangements have been made to enable comprehensive checking also to be carried out, i.e. the input to the loud-speaker can be taken from (*a*) the B amplifier output, or (*b*) the check receiver output.

Test Room.

A small room has been equipped with the necessary apparatus for taking definite measurements of the performance of either of the two transmitters. The intention is that such measurements shall be made once a week, or more often if necessary. The apparatus contained in this room is also used when any alteration has been made to a transmitter, or when the quality is considered below standard for some indefinable reason, or when special experimental work is in progress. The equipment consists of the following:—

(*a*) Cathode-ray oscillograph, used for checking the degree of modulation which the transmitter will take without distortion, and particularly for observing the linearity with a varying degree of modulation.

(*b*) A constant-amplitude variable-frequency valve oscillator with associated amplifiers, for measuring the frequency characteristics of the lines, amplifiers, etc.

(*c*) Comprehensive loud-speaker checking can also be carried out in this room, so that measurements can be taken simultaneously with observations by ear.

Permanent wiring has been installed between both control rooms and the test room, so that these operations can be carried out without disturbing the apparatus used for the routine transmission of programmes. Running from each transmitter to the test room there are also three screened leads which can be coupled to various points in the main transmitter circuit.

Control Room.

Two rooms are provided to house the necessary amplifiers, check receiver and switchgear for controlling and checking the audio-frequency input to the transmitters. Since most of the apparatus has been developed by the B.B.C. specially for use at the Regional transmitting stations, it is proposed to describe it in some detail. Each control room is normally employed in conjunction with the transmitter adjacent to it (see Fig. 2, ground-floor building). Should the necessity arise, however, the control of each programme may be carried out from either position. In order to minimize, as far as possible, trouble from high-frequency currents which might be induced in various parts of the input equipment by the

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transmitters, the control rooms are screened with copper gauze built into the floor, ceiling and walls, the whole being efficiently bonded and earthed.

In each room the actual control position consists of a small mahogany desk upon which are mounted the volume-control potentiometers, monitoring meters, operating keys, etc., all leads to the desk being laid on an overhead cable rack running from the amplifier racks. Upon the latter are mounted the checking apparatus, the " A " and " B " amplifiers, etc., and, in the case of No. 2 control room, the cable terminating and testing equipment. An " A " amplifier is inserted between the " B " amplifier and the microphone when the studio is in use. The output of a " B " amplifier is connected to the input of the transmitter, while its input is derived either from the cable line to Savoy Hill or the output of an " A " amplifier.

There are two battery rooms, one adjoining each control room, that adjacent to No. 1 control room containing the anode-supply batteries and the other the filament batteries with their charging generators.

Cable Circuits.

There are five cable circuits to Savoy Hill, of which four are unloaded and suitable for music transmission, whilst the fifth is lightly loaded and is intended mainly for general communication purposes. The lines are terminated on " U " links in No. 2 control room, from which point the circuits are brought via line fuses to screened repeating coils, the latter being essential in preserving accurate capacity balance between cable and earth.

Distribution of Cable Circuits.—The secondaries of the repeating coils are brought to connection strips at the top of the test rack, at which point the actual distribution of the circuits takes place. Double jumpers connect each music line to two sets of break jacks, upon which terminates the internal wiring from the respective control desks, the main purpose of the jacks being to enable a line to be isolated from either desk should a local fault occur.

On the desks each line is common to two 3-position keys, and a " busy " lamp associated with the line in question lights on each desk as soon as one of the keys in either control room is moved from a central position and a line thus occupied.

The functions of the two keys are:—

(1) To connect a line to one or other of the " B " amplifier inputs (there are two sets of amplifiers in each control room); or

(2) To connect a line either to the break-jack field on the test rack in No. 2 control room or to the desk control telephone common for communication purposes.

With regard to the fifth line, a jumper brings it to a break jack, from which connection is made to a key on No. 2 control desk. This key, when in the normal position, connects the line through to the office, and when in the " off normal " position brings the line to both control desks, where it can be utilized in the same manner as the other lines. A " busy " lamp glows on each desk as soon as this line is occupied.

Communication with Savoy Hill.—During transmission one line is employed on each desk for communication purposes, and should this or one of the music lines fail, two courses are open, one being to utilize the office line (No. 5) and the other to share the remaining control line between both control desks. Obviously, wherever possible, the office line is used for communication and not for transmission purposes, owing to its comparatively bad frequency characteristic.

Each control position is provided with a ringing machine and a hand microphone-telephone set, the latter hanging conveniently at the side of the desk.

Amplifiers.

Each control room is provided with a single " A " and two " B " amplifiers, the former being for use in connection with the local studio equipment as indicated above.

The amplifiers are mounted on the apparatus rack, their feed meters and the microphone current meter being fitted in a panel above them, whilst the " B " amplifier control potentiometers are placed in a convenient position on the control desks.

For test purposes both inputs and outputs of all amplifiers are terminated on break jacks, a strip of which is mounted on each amplifier rack. The amplifier filament circuits are completed by means of keys, placed within reach of the control engineer, on the rack just above the amplifiers themselves. H.T. and L.T. fuses are connected in each circuit.

Checking Receiver.

Each control room is provided with complete equipment for checking by wireless reception.

The performance for which the receiver was designed is as follows:—

Selectivity.—The receiver is to be capable of completely cutting out signals of equal strength separated by 50 kilocycles per sec. from the frequency to which the receiver is tuned, the selectivity to be such that there is no audible trace of the interfering signal.

Audio-Frequency Response.—The overall frequency-response characteristic of the receiver should be flat from 20 to 10 000 cycles per sec. within $\pm$ 0·5 decibel.

Monitoring Arrangements.

(*a*) *Visual.*—Three meters are available for visual monitoring at each control desk, these being the modulation meter, the slide-back galvanometer and the speech-choke voltmeter. The latter is operated in series with that joined across the speech choke in the anode circuit of the modulated magnifier stage of the transmitter, and is calibrated accordingly. As a general rule this voltmeter is used in preference to the other meters, as it gives the engineer more direct check over the performance of the transmitter.

Slide-Backs.—Each control room is fitted with a standard slide-back unit, incorporating two separate instruments. These are each connected permanently across one of the " B " amplifiers, separate meters being mounted on the control desks.

A slide-back unit consists of a triode valve with a negative grid-potential, the low-frequency input to which can be adjusted accurately by means of a potenti-

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ometer. A sensitive d.c. milliammeter is connected in the plate circuit to indicate when anode current is flowing. The grid and anode potentials of the slide-back valve are set, and the input potentiometer is so adjusted that anode current flows only when the transmitter is receiving its maximum audio-frequency input.

(*b*) *Aural.*—Loud-speakers, operated by the comprehensive checking amplifier, are available for programme checking in both the quality rooms and the control rooms, although routine checking is carried out by means of headphones only in the latter. Jacks for the headphones are fitted at each transmitter control desk and in both control rooms, so that, if required, both programmes may be heard by the engineer on duty.

In order that telephonic communication may be carried out without interruption when the loud-speaker is in use, a pair of contacts from the desk telephone speak-ring key are arranged to operate a relay which breaks the loud-speaker circuit when the key is in the " speak " position.

The loud-speaker amplifier is not used for routine monitoring, but feeds loud-speakers in the test room, engineer-in-charge's office, mess room and waiting room, etc. Two lines are run to each of these points, one from each checking equipment, and a pair of jacks is provided to enable a choice of programmes to be made.

Test Rack.—As previously stated, any of the five circuits from Savoy Hill may be brought to the break-jack field on the test rack in No. 2 control room, by throwing a key on one or other of the control desks. Similarly, all amplifiers, etc., are terminated on break jacks, and they may thus be utilized for test or tested themselves by connecting them to the break-jack field by means of double-ended cords. Tie lines are brought from the jack field in No. 1 control room to that in No. 2, in order that the apparatus in the former may be available for use in the test bay.

Break Jacks.—All break jacks employed on the apparatus are of the double type, i.e. they consist of two 3-point jacks (tip, inner and sleeve) mounted one above the other.

The plugs used in connection with these jacks are serrated on one side of the grip, in order that polarity may be traced when they are used on a double-ended cord.

Apparatus for Maintaining Accuracy of Radiated Carrier Frequency.

A room has been set apart for apparatus for maintaining a specially high degree of constancy of the carrier wave of one of the transmitters. This has been done to cover the possibility of working one of the transmitters on the same radio frequency as one or more other B.B.C. stations. The system used at present for several of the smaller transmitters working on a single frequency depends on a thermostatically controlled tuning-fork, which produces a constant frequency of the order of 1 000 cycles per sec. This frequency is then multiplied by frequency-doubling amplifiers until the required radio frequency is obtained.* It is possible that such apparatus may also be installed at Brookmans Park for the purpose of emitting a standard frequency of declared accuracy, which would be of the order of 5 parts in a million, assuming the tuning-fork method of drive. Working the station as it is at present, i.e. relying only on the master oscillator, which is a D.E.T.2 valve working at a power level of about 200 watts, provided with filament heating from a battery and a separate motor-generator for the anode supply, the accuracy is of the order of 300 cycles over a period of two or three days for a carrier-wave frequency of 1 148 kilocycles per sec.

* P. P. Eckersley and A. B. Howe: " The Operation of Several Broadcasting Stations on the same Wavelength," *Journal I.E.E.*, 1929, vol. 67, p. 772; also *Proceedings of the Wireless Section*, 1929, vol. 4, p. 81.

Studio.

A small studio has been provided in order to enable tests to be carried out locally, and for use in emergency. The walls are treated in the normal manner in order to give satisfactory results acoustically. Gramophone apparatus has been installed in addition to the usual microphone equipment.

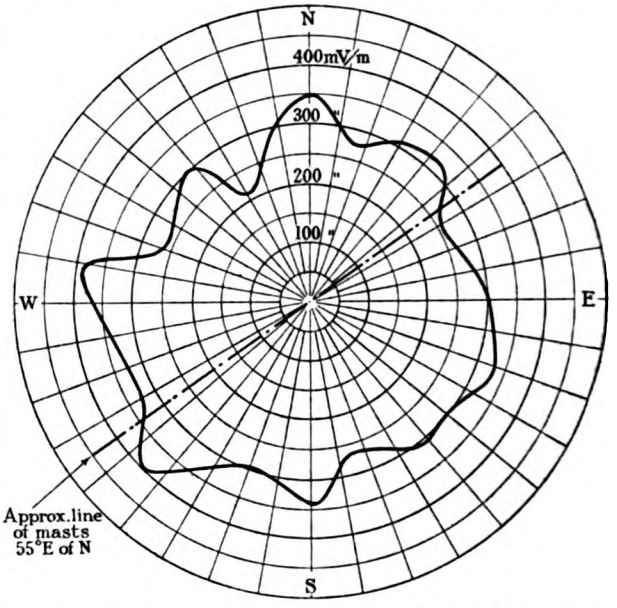

Fig. 16.—Polar diagram of 356-m transmitter, with approximately 30 kW in the aerial.

Alarm System.

In order to summon the engineer-in-charge of a shift to any part of the building where assistance may be required, a system of coloured signal lights has been installed in several positions, such as the transmitter hall, power house, control rooms, etc. One particular colour is associated with each point, for instance a red lamp may indicate the power house, a blue lamp the transmitter hall, etc. When assistance is required, the attendant presses a button on the wall and a lamp having the colour allotted to that particular room lights up in all the other rooms where the shift engineer is likely to be, a bell ringing in each room at the same time in order to attract his attention to the lamp. The engineer can then proceed straight to the room indicated. This system, together with the internal telephone system, was installed by Ericsson Telephones, Ltd.

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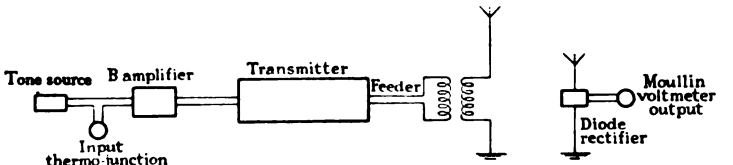
GENERAL PERFORMANCE OF THE STATION.

The effect on the listening public of the change-over from a low-power station situated in the centre of London, to a comparatively high-power station approximately 15 miles due north of London, has already been discussed in an earlier section of the paper. This concerns, of course, the transmitter which is tuned to a frequency of 842 kilocycles per sec. (356·3 m). The input to the high-frequency feeder lines in this case is approximately 30 kW, and the polar diagram of field strength for this power, when taken at a distance of 5 km from the station, is shown in Fig. 16. It must be realized that polar diagrams express a mean of field strength, and that idiosyncrasies of particular localities produce large variations from this mean

Fig. 17 gives the overall frequency-response curve including the " B " amplifier, transmitter and aerial. The response of the transmitter to varying amplitudes of audio-frequency input at a given frequency is sensibly flat up to 100 per cent modulation. The frequency of 842 kilocycles per sec. (356·3 m) is transmitted on the north masts, which are insulated at the base. The effect of the masts under these conditions is negligible at this frequency. As will be seen from Fig. 1, the two aerials are parallel and the distance between the line of each aerial is 900 ft. At this distance the effect of the masts of one transmitter on the other is also practically negligible, whether they are earthed or unearthed. This fact is apparently due to the distance alone, because the effect on either transmitter of earthing or unearthing the masts of the other transmitter is very small, while the effect of earthing or unearthing the masts in use for either transmitter is very considerable. In the case of the 1 148 kilocycle-per-sec. (261·4 m) transmitter the polar diagram can be varied by different methods of earthing the masts, and work on this problem is still proceeding, but the conditions of working at the time of writing the paper are shown in Fig. 18. The power supplied to the aerial system has since been raised to 45 kW.



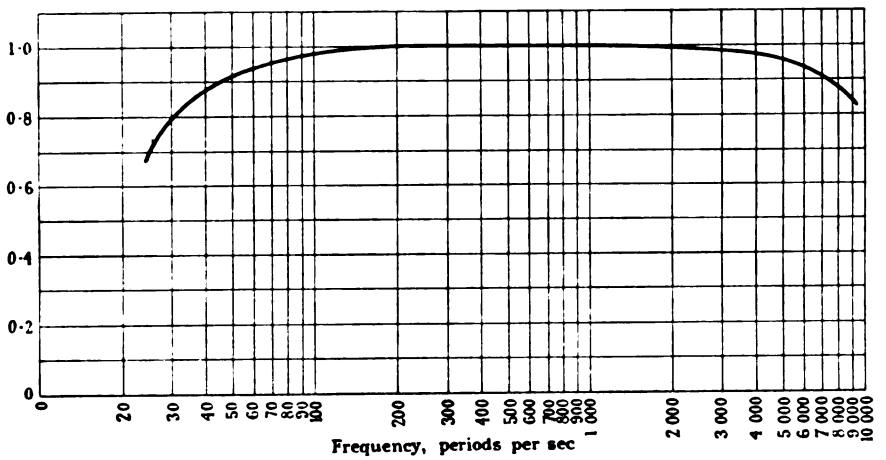


FIG. 17.—Overall response curve of 356-m transmission.

Observations of variations of frequency of the carrier wave due to modulation (frequency modulation), made both in the static and dynamic conditions, indicate that

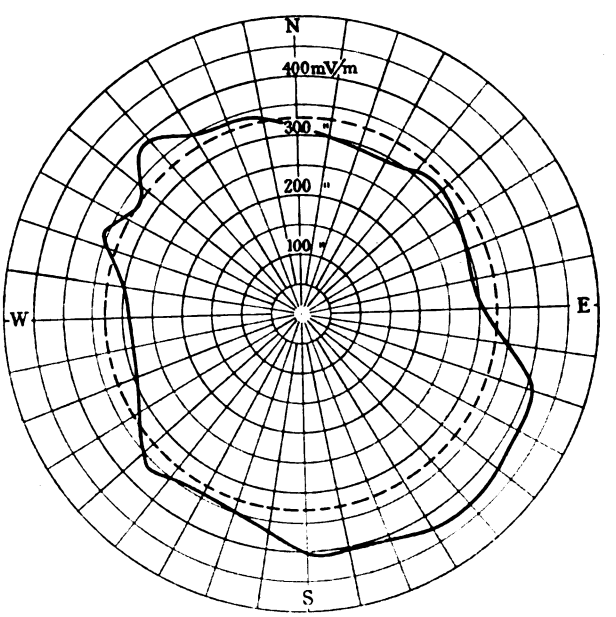


FIG. 18.—Polar diagram of 261-m transmitter with approximately 30 kW in the aerial.

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this was so small that it was extremely difficult to obtain any definite figure.

No difficulties of an interesting nature were encountered when testing the station in the preliminary stages, although the transmitters had not been tested as complete units at the makers' works. Certain minor defects and omissions of little importance showed themselves, but in no case was any major alteration made.

It had been anticipated during the design stage that some difficulty would be experienced from "cross-talk" between the two programmes. There was, of course, the possibility that there might be some mutual high-frequency effect between the two transmitters themselves and between the feeder lines and aerial systems, but this effect was found to be absent to a rather remarkable extent. When the transmission from one aerial was unmodulated and a pair of telephones was connected across the check receiver at a point where the power level was suitable for a loud-speaker, no trace of the modulated radiation from the other transmitter could be heard. This was carried out with the control-room apparatus disconnected from the transmitter under test. This test is, of course, extremely drastic because no practical listening could be carried out under equivalent conditions. However, when one transmitter was radiating a programme and the other was emitting an unmodulated carrier, but with the input apparatus fully energized, distinct cross-talk was audible. This was found to be due to the fact that two earths had been provided for each control room, on the principle of a "silent" and "noisy" earth. The filaments of the amplifier valves were connected to the "silent" earth, and the amplifier screening-boxes and the gauze screen enclosing the whole control room were connected to the "noisy" earth. This scheme of connections resulted in a pick-up of the modulated high frequency from the other transmitter, which found its way through the amplifier to the transmitter, becoming sufficiently rectified on the way to produce an audio-frequency component of the unwanted programme. This was rectified very simply by using a common earth for all apparatus in the control room, no other effects being introduced by this altered method of connection.

Conclusion.

The authors' thanks are due first of all to the Governors of the British Broadcasting Corporation, for permission to publish this paper, and to the following members of the B.B.C. staff, who have made considerable contributions to the design and construction of the station itself: Mr. H. Bishop, the assistant chief engineer; Mr. H. L. Kirke, head of the research department, and his assistant, Mr. T. C. Macnamara; Mr. M. T. Tudsbery, the civil engineer; and Mr. B. N. MacLarty, head of the installation section. Finally, thanks are due to Mr. L. W. Hayes, for assistance in correcting diagrams and proofs, and to Mr. J. E. Calverley, of the English Electric Co., for supplying information in connection with the large high-tension motor-generators.

Discussion before the Wireless Section, 7th May, 1930.

Col. A. S. Angwin: I am very interested in the essential differences between this, the most modern wireless station for broadcasting, and the general practice in radio-telephony stations for commercial point-to-point working. One of the most striking features of the new station is the finality which presents itself in the lay-out. It is obvious that the authors are not contemplating turning their twin transmitters into triple transmitters at any very early date; neither are they likely to embark on a competition with European broadcasting stations in augmenting the power of the station. The ground space is adequately filled and they have presented a complete lay-out sufficient for the immediate problem, without provision for expansion in the future. I think that this limit to the possible expansion must have largely influenced the authors in deciding whether to take power from outside or to use a self-contained system. For a station of this power (500 kW) and with the very good running load-factor, the conditions are undoubtedly as favourable as they could be for a self-contained power system, but the statement actually made by the authors in dealing with the question of their choice: "Moreover, it is possible to generate locally at approximately the same cost as, or in some cases a considerably lower cost than, that at which power can be bought," merits some attention. The authors' contention does not agree with our experience for power supply to commercial radio stations. From the remarks which Captain Eckersley made in presenting the paper, I gather that cost was not the main consideration. If power of the order of 500 kW can be generated locally more economically than it can be bought in such a location as London, close to a power supply, it does not reflect favourably on the scheme for the adequate supply of power at a reasonable cost which is at present covering the country. With regard to the factor of breakdown, the authors state: "It is interesting to note that 'shut-downs' of considerable duration at broadcasting stations have frequently been due to failure of the public supply mains." This seems to be a grave reflection on the reliability of a public supply. I think that, in spite of the fact that provision has been made for a complete spare Diesel engine and generator, the other sources of trouble common to the motor-generator plant are in themselves as likely to cause interruption as anything that might normally be expected from a well-regulated outside power system. It is interesting to note that the designers have stood their ground and have provided for a high-tension d.c. supply rather than a rectified a.c. supply. It is obvious, however, that consideration of the trouble which they have experienced was ever present with them when the station was designed, and precautionary measures have been necessary, such as the addition of chokes and resistances in the anode supply to the valves. The device of adding a resistance at each anode feed is, it seems to me, essentially a compromise. It is clear that much work remains to be done on the problem of connecting a high-tension d.c. supply

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from a motor-generator direct to water-cooled valves such as those at present in use. The insertion of a resistance of the high order of 500 ohms directly in the feed circuit seems rather like throwing away the efficiency which is striven for by the use of high-tension d.c. generators. I have made a small calculation which shows that a resistance of 500 ohms dissipates about 0·5 kW of energy in the anode supply to each valve. Taking the cost of power as 1d. per unit, this is ½d. per hour per valve. I estimate that the valves themselves are costing 3½d. per valve per hour to run. From this, it seems that unless an increase in the life of the valve of the order of 14 per cent is secured it is a very uneconomic means of providing for a lengthening of valve life. I presume that the station has not been running long enough for any figures to be available as to what order of life is being obtained from the valves, but it is of importance to consider how much one can afford to waste by adding resistance in the anode circuit of a valve in order to improve the life of the latter. There are other considerations which have doubtless carried weight with the designers, such as greater security from temporary breakdown. It is stated on page 212 that the present accuracy of frequency is of the order of 300 cycles per sec. in 1 148 kilocycles per sec. I congratulate the designers on having at once obtained a tolerance which was suggested at the last C.C.I.R. Conference. This is done by the use of a master oscillator. The C.C.I.R. has also recommended that within a short time this limit of tolerance of a broadcasting station should be brought down to 50 cycles. It would be interesting if the authors could give some indication of how they hope to achieve this, whether by means of a master oscillator or by the use of a tuning fork. A photograph of the high-frequency transmission line has been shown on the screen and it is briefly mentioned in the paper. Some details of the actual type of line employed would be of great interest. I should also welcome further information regarding the methods which the authors have adopted for altering the shape of the polar diagram (Fig. 18). I infer that there is more involved than merely insulating or earthing the masts; presumably something in the nature of a tuned circuit is interposed between mast and earth.

Mr. J. E. Calverley: I should like to refer to the high-tension machines used in the station. The firm with which I am associated has been manufacturing high-tension machines for over 20 years. The first were built about 1907 and were of 50-kW capacity at 5 000 volts. They had air-insulated commutators, and insulated foundations and couplings. The present machines have been developed in the course of time from those original sets. In the meantime, about 20 different sets have been built for voltages of 10 000 and over. The first machines were used for telegraphy, and four were connected in series to give 20 000 volts. To prevent large variations of load on the generators a battery of 10 000 cells was used in parallel. There was an air of experiment throughout, but in the present station there is apparently nothing experimental at all. The machines are normal, the lay-out is normal, the cables are taken in an orderly manner in ducts and there is little to distinguish the station from one running at 500 volts d.c. This very real progress is the more evident by reason of its unobtrusiveness. The first point upon which a designer would seize when designing the high-tension machine is the apparent advantage offered by 2-pole windings with wide spacing from brush to brush. The machines adopted by the authors are of the 4-pole type, and only when one is building a 2-pole machine is its weakness apparent. The insulation to earth and between conductors of the long end-windings is a difficult matter with 2-pole machines. In d.c. machines, where the slot pitch is limited by commutating conditions, the maximum thickness of insulation is probably in the neighbourhood of 0·1 in., and with 10 000 volts to earth the stress on the slot insulation is then 100 000 volts per inch, which is too high for continuous operation. For this reason it is usual to insulate the machines from earth, but this has its disadvantages, being expensive and dangerous and necessitating insulating couplings. It is quite a definite step forward to take such precautions that the insulation stresses are graded throughout from positive to earth in such a way that the stresses are normal and, further, that they are so controlled that no undue stress can arise in any part, even in the case of accidents. I should be interested to know how extensive a development of high-tension machines will be required during the next few years. What outputs and voltages will be required? It is difficult to design machines for wireless experts owing to their constant new demands, but the limit of size for this class of machine does not seem to be in sight. It will be safe to make much larger machines than have been built up to the present. In view of the general popularity of rectifiers, I think that much credit is due to the engineers of the B.B.C. for adopting d.c. machines with their higher efficiencies and greater robustness.

Mr. E. H. Shaughnessy: I am interested to see that the B.B.C. have taken up the closed water-cooled system. I believe that the first closed system was installed at the Rugby radio station, where we used distilled water. I am also glad to note that the authors have adopted the practice which we favoured at Rugby of using d.c. machines for the high-tension supply to the valves. At Rugby we have d.c. machines made by the British Thomson-Houston Co., each unit of which is of 250 kW, has only one commutator and gives 3 000 volts, so that with two machines we develop a pressure of 6 000 to 7 000 volts for the high-tension supply. These have now been running satisfactorily for some years. I congratulate the engineers of the B.B.C. on the performance of the Brookmans Park stations. The reception that we obtain in South London is quite as good from the new station as it was from the old 2LO. On Sunday last, however, I was listening to some very quiet music, by Wagner I think, on the National programme, when I heard a religious service which was being broadcast on the same wavelength, and which was also being radiated from the second Brookmans Park transmitter. On testing my receiver I found that nothing was wrong with it. I should be interested to know whether this overhearing was due to the land lines, some coupling in the station at Brookmans Park, or the arrangement at Savoy Hill.

Mr. P. K. Turner: The new station has a good audio-frequency response curve—far better than that of

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any other broadcasting station I have seen. In connection with the possibility of good reception from the station, I notice that it is so designed that it can be modulated up to 100 per cent without distortion. I do not know quite what this means. Can the authors give figures showing what they regard as the mean and the peak modulation percentages of the transmitter? I would ask them not to make these figures too high, as although there is everything to be said, from certain aspects, for pushing up the modulation, it rapidly becomes almost impossible to rectify this in the receiver without undue distortion. It is quite easy to rectify matter which is modulated 60 per cent, but with modulations of 80 or 90 per cent it becomes increasingly difficult to rectify without bad wave-form distortion.

Captain P. P. Eckersley and **Mr. N. Ashbridge** (*in reply*): Colonel Angwin questions *inter alia* whether more allowance should have been made in the general design of the station for future extensions. The existing plant allows for an output from each transmitter of 50 kW, although it was intended to operate them with an output of 30 kW. The latter power was actually used when the station was opened, but in the case of the transmitter using a frequency of 1 148 kilocycles per sec. (261·3 m) the power has since been raised to 45 kW in order to minimize, as far as is possible, the effects of local shielding, which are somewhat severe in certain districts on this comparatively high frequency. We doubt whether a large increase in power over and above 50 kW to the aerial is justifiable economically, for transmitters using these frequencies when considered in relation to the increased performance obtained. With regard to the question of a large-scale extension, such as the installation of a complete additional transmitter, sufficient land is in fact available.

Colonel Angwin has expressed surprise that it should be possible to generate electric power locally at a price comparable with that charged by a public supply company. In this connection it has to be remembered that the nature of the load is very advantageous, and with Diesel engines operating at their optimum load the cost of generation should be low. In comparing the two possible methods of providing power, account must be taken, firstly, of the fact that the provision of some proportion of stand-by generating plant was considered to be essential on account of the prime necessity of continuous service; and, secondly, of the cost of cables and substation plant. With regard to the question of the relative reliability of a public supply as compared with local generation, it is not really a comparison of the reliability of one generating station with another, because in our experience the failures of public supplies have almost invariably been due to the mains and not to the generating stations. It is true that failures elsewhere in the station are likely to be more frequent than supply failures, but the breakdown of a single component cannot be compared with the complete failure of supply, which entirely puts out of action both transmitters, together with all auxiliary plant such as motor-generator sets, water pumps, etc. Apart, therefore, from the duration of the actual mains' failure, some additional time is lost in restarting the whole station.

We agree that the insertion of a 500-ohm resistance in the anode supply lead to each main power valve is a compromise, but it is hoped that with improved valves this may not be necessary in future. It is done at present more to minimize the effect of an unstable valve than to lengthen the life of the valves, although to an extent the latter is an additional advantage.

Colonel Angwin inquires what steps could be taken to increase the constancy of the carrier frequency, should this become necessary. It is probable that a thermostatically controlled tuning-fork will be used, but the possibility of a "crystal" drive will not be overlooked. The whole question is actually bound up with the ultimate object in view; for instance, whether constancy is most desired, and whether absolute accuracy is necessary in addition, and so on. It should be explained that the stability of carrier frequency of 300 cycles per sec. in 1 148 kilocycles per sec. mentioned in the paper is obtained without any day-to-day readjustment of the drive circuit of the transmitter. This figure is reduced in practice to a variation of about 100 cycles per sec. by an occasional slight re-adjustment of the drive to take account of temperature variations, the amount of re-adjustment necessary being indicated by a wavemeter of the type designed by the Union Internationale de Radiodiffusion.

The feeder lines are composed of 19/·044 phosphor-bronze wire, supported on tubular posts 13 ft. high. The insulators are of standard pattern, and were tested at 7 000 volts, 1 200 kilocycles per sec., for 4 hours. There is nothing unusual in the construction of these feeder lines, and similar feeders have been erected elsewhere in the B.B.C. system. Data as to how the shape of the polar diagram was corrected are not yet available for publication. It is probable, however, that they will be contained in a future paper.

Mr. Calverley inquires what outputs and voltages are likely to be required for high-tension d.c. motor-generator sets in the future. We do not consider that it is possible at the moment to estimate the future requirements of designers of broadcasting stations. There are, of course, alternative methods of producing the necessary high-tension direct current for wireless transmitters, and it is difficult to say which will ultimately gain the most favour for all types of radio stations. Speaking of broadcasting, it would appear that with a self-contained plant high-tension d.c. generation is most economical and flexible. If, however, a.c. mains supply is used, it is possible that the mercury-arc rectifier would show substantial economies over any other system. From the point of view, however, of the particular station under discussion, the mercury-arc rectifier is not considered to be superior to d.c. generation, and, therefore, there seems to be an important application for the method adopted by the Corporation.

Mr. Shaughnessy refers to cross-talk which occurred on a certain occasion between the two programmes radiated from Brookmans Park. It is true that cross-talk has occurred on occasions, but in most instances this has not been due to the transmitting station itself. There has been cross-talk between external telephone lines as well as in the control room at Savoy Hill, but in both cases steps are being taken by the Post Office and the B.B.C. respectively to reduce the number of occasions on which it occurs.

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In reply to Mr. Turner, whilst it is true that the paper states that the transmitters are capable of being modulated linearly to the extent of 100 per cent it does not mean that the transmitters are necessarily run under these conditions. It is not possible to give a definite figure for the average modulation of a transmitter, because this varies with every type of programme. The maximum peak modulation normally reaches about 80 per cent, and the average is very roughly in the neighbourhood of 20 per cent, but, as already mentioned, the latter figure cannot be stated with accuracy.

DISCUSSION ON

"THE DEVELOPMENT OF A PRECISION AMMETER FOR VERY HIGH FREQUENCIES." *

Mr. W. F. Dunton (*communicated*): In Section (2) of the paper the author quotes a well-known but incorrect formula for the electromagnetic force between parallel conductors. The correct formula, first published by me in 1927, is now widely known among electrical engineers on the Continent. If two straight parallel conductors, each s cm long, have a direct current of I amperes uniformly distributed over their circular cross-sections, and if the distance between their axes is a cm, the force (in pounds) with which these conductors tend to attract or repel each other (according to whether the currents are similarly or oppositely directed) is not $4{\cdot}5\,(I/10\,000)^2 s/a$ but $4{\cdot}5(I/10\,000)^2\{\sqrt{[(s/a)^2+1]}-1\}$. Mr. Moullin says that the force is given by the first of these expressions when the conductors are "long"; but he does not give any idea as to the length which he has in mind, and he does not give any formula for other lengths. The second expression will, however, give the actual force in all cases and, to an engineer with a slide rule, is almost as convenient as the first. In a circuit breaker $(s/a)^2$ is frequently less than $\frac{1}{4}$, and in such cases the last factor in the second expression is approximately $\frac{1}{2}(s/a)^2$. The results given by the first expression are then $2a/s$ times what they ought to be. But a/s is frequently greater than 3, and the results obtained with the first expression are then more than six times what they ought to be, no matter how long the conductors may be. On the other hand, if $(s/a)^2$ is much greater than 1, the last factor in the second expression becomes approximately $(s/a-1)$. Therefore, even where the length of each conductor is many times the diameter while the distance between their axes is only a few times the diameter of either conductor, the first expression would be improved by substituting $(s/a-1)$ for s/a, a change which obviously does not make the formula less convenient. A detailed proof of the formula will be found in the *Journal of Scientific Instruments* dated November 1927, and in the *Revue Générale de l'Électricité* dated 12th May, 1928.

* Paper by Mr. E. B. Moullin (see page 133).

Mr. E. B. Moullin (*in reply*): The expression $F = 2I^2D$ per unit length is of course the limiting value which is approached when the length of the conductors is very great compared with their distance apart. In lines 20 and 24 of page 137 the length is given as 10 cm, and at the bottom of the second column of that page, and at the top of the second column of page 139, the value of D is found to be 1·11 cm for one instrument and 0·87 cm for the other. The length of the conductors is therefore about 10 times the distance between their centres. The approximate form of Mr. Dunton's formula, namely $F = 2I^2(s/D-1)$, is a very useful and convenient abbreviation of the correct expression for the force between wires of finite length. It shows that the force in these instruments is about 10 per cent less than is given by $F = 2I^2/D$ per unit length. This deficit produces no real error, because the instrument is calibrated from another ammeter and not from the measured value of the force. It has also a small effect on the correction factors, but the effect is of the second order of magnitude. The total correction factor for proximity is about 4 per cent; over a small portion of the length this factor will be perhaps a little less than 4 per cent and so the total factor will be reduced by a very small amount, certainly less than 10 per cent. But the factor itself is small, so the correction to the correction is negligible and quite outside the accuracy of reading. The effect will be more pronounced in the correction factor for the screen tube and this is implicit in the statement near the bottom of page 139. The screen factor can, however, be measured and we need not rely on calculated values; thus it is found to be 6·6 per cent when the calculated value was 8 per cent. The effect of the finite length of the cylinders has been fully appreciated, but it has no measurable effect on the accuracy of the instrument, because the calibration is made from another ammeter. The "effect of irregularities at the ends of the system" was also dealt with in section (8) on page 67 of vol. 121 (1928) of the *Proceedings of the Royal Society*.

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GENERAL FORMULÆ FOR THE RADIATION DISTRIBUTION OF ANTENNA SYSTEMS.*

By R. M. WILMOTTE, M.A., Associate Member.

[From the National Physical Laboratory.]

(*Paper first received 12th September, 1929, and in final form 27th February, 1930.*)

SUMMARY.

Formulæ are given for the distribution of radiation in different directions. The paper is divided into the following sections:—

(1) Introduction.
(2) Antenna elements.
(3) Polarization.
(4) Current distribution.
(5) Effect of image in the ground.
(6) A straight wire inclined at any angle.
(7) A system of similar antennæ spaced in any given manner.
(8) A system of antennæ uniformly distributed around the circumference of a circle.

A number of special cases are deduced from these general formulæ, and curves are given to facilitate computation.

(1) INTRODUCTION.

It was considered that it might be of value to combine in one paper some general formulæ for the distribution of radiation from antennæ in various directions as far as present theories permit. These theories are not yet rigorously established and much experimental confirmation is still required. Many special cases of radiation from antenna systems have already been worked out by a number of authors. These, as far as they are known to the present author, are given in the Bibliography [items (1) to (10)]. Most of the formulæ are special cases of those given below, but special credit should be given to Abraham,† who initiated the principle of representing an antenna by a series of doublets, and to Pierce,‡ who first extended it.

In employing the formulæ given below it must be remembered that all the conducting currents must be taken into account. The presence of masts, stays and even underground conductors may affect the radiation distribution of an antenna system to a very considerable extent, owing to the currents induced in these conductors from the main antenna system. The extent to which this may occur is strikingly shown when an attempt is made to screen a portion of an antenna. Smith-Rose and Barfield § published some interesting results on this subject. They found that, on long waves, practically no effect on the signal strength was produced when the vertical portion of an inverted "L" antenna was surrounded by a metal screen. Evidently the currents induced in the screen induced in the horizontal portion an e.m.f. roughly equal to that which would have been induced in the vertical portion of the antenna had the screen not been there. Removal of the horizontal portion reduced the signal practically to zero. If such is the case it is to be expected that considerable alterations in the radiation distribution may be expected to be produced by any neighbouring conductor. It is probable that the result obtained by many workers, that a horizontally polarized ray on the ground is always rapidly rotated into a vertically polarized ray, is largely due to this effect. The horizontally polarized ray is so rapidly attenuated compared with the vertically polarized ray that its strength at any reasonable distance is negligible and the signal received is produced by vertical conductors in the neighbourhood of the transmitting system. This may also be produced by groups of horizontal and vertical conductors sufficiently close to the transmitter for the currents induced in the horizontal members to induce sufficient current in the vertical members to enable a definite signal to be received on a vertical antenna at a reasonable distance.

Many of the formulæ given below are very cumbersome. For computation purposes the curves given in Figs. 4 to 9 and 12 and 13 will be found useful.

Before embarking on the analysis it may be of interest to consider what is exactly meant by "radiation distribution." The property of the antenna system which this term is intended to convey quantitatively is the variation of field strength in different directions when the attenuation is neglected. In this connection there are two terms which are commonly used but have no rigorous meaning and largely overlap one another. They are "polar diagram" and "effective height." Effective height is a factor which, when suitably applied in Hertz's radiation formulæ, gives the field strength. A single number as a representation of this property is meaningless when applied to a directive system. Often there is no ambiguity in its meaning, but a combination of this with the polar diagram, which itself purports to give the variation of field strength with direction, would be more rigorously correct. This then would be a function of angular directions which, when applied, as explained below, in Hertz's equations, gives the signal strength in various directions when the attenuation is neglected. This factor may for convenience be called the "radiation distribution of the antenna system."

This function should give not only the strength of the field but also its direction and phase. A convenient

* Reprinted from *Journal I.E.E.*, 1930, vol. 68, p. 1171.
† See Bibliography, (1). ‡ *Ibid.*, (2). § *Ibid.*, (11).

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method of expressing this is by means of two functions, one for each of the components of the electric field, these two components being perpendicular to the direction of propagation. The radiation distribution will, therefore, consist of two functions containing direction (relative to the antenna) and phase as variables. The field at a point P having polar co-ordinates (r, θ, ϕ), where θ is the vertical angle, ϕ the horizontal angle (see Fig. 1) and r the distance, is large compared with the dimensions of the antenna system, is given by

$$E_\theta = \frac{60\pi I}{\lambda r}\left[f_\theta \cos \omega\left\{t - (r'/c)\right\} - f'_\theta \sin \omega\left\{t - (r'/c)\right\}\right] \text{ volts per metre} \quad . \quad (1)$$

$$E_\phi = \frac{60\pi I}{\lambda r}\left[f_\phi \cos \omega\left\{t - (r'/c)\right\} - f'_\phi \sin \omega\left\{t - (r'/c)\right\}\right] \text{ volts per metre} \quad . \quad (2)$$

where E_θ is the longitudinal and E_ϕ the latitudinal component of the electric field, λ (metres) is the wavelength, c (metres per sec.) the velocity of light, ω is

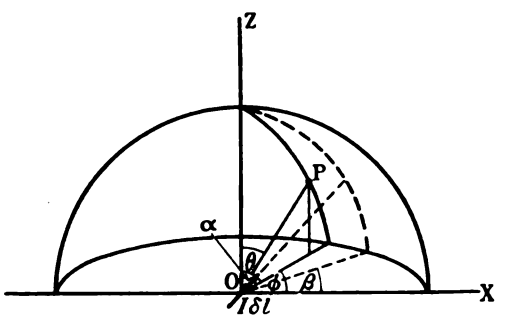

FIG. 1.

$2\pi \times$ frequency, and r' (metres) is the equivalent distance, as explained in a previous paper by McPetrie and the author.*

The radial component is not given, as the distances are usually so large as to make this negligibly small. These equations give the field strength in any direction when there flows in the antenna current of value $I \sin \omega t$ amperes. As this current is not the same in all parts of the antenna system it is essential to state the point where it is measured. The functions f_θ, f'_θ, f_ϕ, f'_ϕ are all functions of direction and represent the radiation distribution. It will be convenient to write the radiation distribution in the form

$$(f_\theta + jf'_\theta); \quad (f_\phi + jf'_\phi).$$

(2) RADIATION DISTRIBUTION OF ANTENNA ELEMENTS.

(a) *Straight Element of Wire.*

The variation of field strength around an element of straight wire is sinusoidal and symmetrical about the axis of the wire. It is, therefore, possible to resolve the element of wire into components as though it were a vector. Consider an element of wire in free space of length δl metres situated at the origin in the plane $\phi = 0$ and making an angle α with the axis OZ (Fig. 1). Suppose it carries a current $I \sin \omega t$ amperes. The field in the plane $\phi = 0$ will be given by

$$E_\theta = \frac{60\pi}{\lambda r} I\delta l \sin(\theta - \alpha) \cos \omega\left\{t - (r'/c)\right\} \text{ volts per metre}$$

and $$E_\phi = 0.$$

These formulæ do not take account of the image in the ground.

In order to find the field at the point P (Fig. 1) having co-ordinates (r, θ, ϕ) from a similar wire element but in a plane making an angle β with the plane XOZ, the element can be resolved in and perpendicular to the plane POZ. It can then be readily shown that

$$f_\theta + jf'_\theta = [\cos \alpha \sin \theta - \sin \alpha \cos \theta \cos(\phi - \beta)]\delta l \quad (3)$$

and $$f_\phi + jf'_\phi = -\sin \alpha \sin(\phi - \beta)\delta l \quad . \quad . \quad (4)$$

(b) *Element of Coil.*

A straight element of wire is not the only form of antenna which can be represented by a line and treated as a vector. The field from a coil has also cylindrical symmetry and varies sinusoidally with the angular direction relative to that axis. It is, therefore, possible to represent a small coil by means of a vector perpendicular to the plane of the coil and passing through its centre. It is necessary to use some arbitrary convention regarding the direction of this vector, and the author suggests that, by analogy with other conventions of signs, the right-handed screw rule should be used as explained by Fig. 2. There is an important difference between the vector representation for a straight wire and for a coil. From a straight wire the direction of the electric field is in the plane containing the vector representing the element of wire and the direction of propagation considered, while for a coil it is perpendicular to this plane. Such a representative vector can be resolved into components as desired and, by reasoning similar to that followed in the case of elements of straight wire, the following results can be obtained for a coil represented by the arrow in Fig. 1.

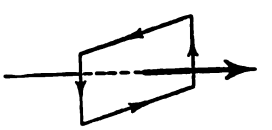

FIG. 2.

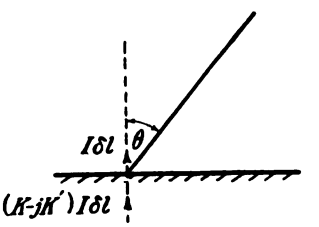

FIG. 3.

If δA represents the area-turns of the coil in square metres, and the current is $I \sin \omega t$ amperes, then the field in the plane $\phi = 0$ will be given by

$$E_\theta = 0$$

$$E_\phi = -\frac{120\pi^2}{\lambda^2 r} I\delta A \sin(\theta - \alpha) \sin \omega\left(t - \frac{r'}{c}\right) \text{ volts per metre}$$

As before, these formulæ do not take account of the image in the ground.

* See Bibliography, (12).

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Hence, from the field at any point P, we shall have

$$f_\theta + jf'_\theta = j\frac{2\pi}{\lambda} \sin \alpha \sin (\phi - \beta)\delta A \quad . \quad . \quad (5)$$

and

$$f_\phi + jf'_\phi = j\frac{2\pi}{\lambda}[\cos \alpha \sin \theta - \sin \alpha \cos \theta \cos (\phi - \beta)]\delta A \quad (6)$$

It follows that any result which applies to a given distribution of vectors representing elements of straight wires will also apply to an identical distribution of vectors representing elementary coils, if we write $-E_\theta$ for E_ϕ, E_ϕ for E_θ, $2\pi I\delta A/\lambda$ for $I\delta l$ and advance the field by 90°.

(i) Vertical wire:—

$$f_\theta + jf'_\theta = \sin \theta \delta l; \quad f_\phi + jf'_\phi = 0 \quad . \quad . \quad (7)$$

(ii) Horizontal wire parallel to X axis:—

$$f_\theta + jf'_\theta = -\cos \theta \cos \phi \delta l; \quad f_\phi + jf'_\phi = -\sin \phi \delta l \quad (8)$$

(iii) Horizontal wire parallel to Y axis:—

$$f_\theta + jf'_\theta = -\cos \theta \sin \phi \delta l; \quad f_\phi + jf'_\phi = \cos \phi \delta l \quad (9)$$

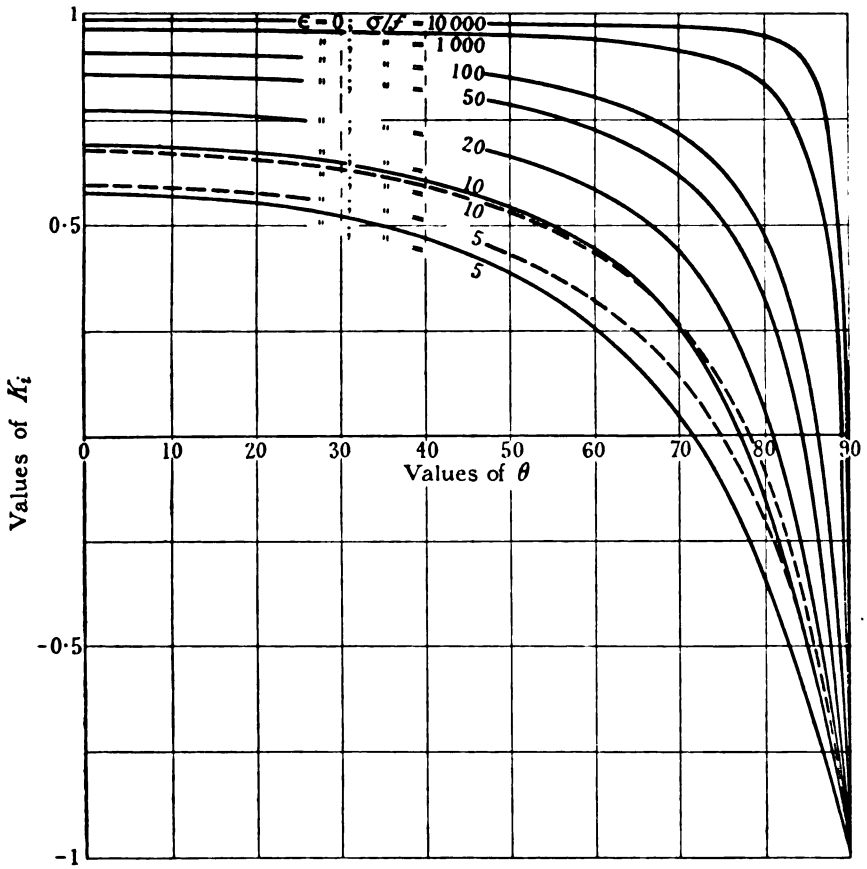

FIG. 4.

$$K_i = \frac{(\varepsilon^2 + 4\sigma^2/f^2)\cos^2\theta - (c^2 + d^2)}{(\varepsilon^2 + 4\sigma^2/f^2)\cos^2\theta + (c^2 + d^2) + 2\cos\theta(\varepsilon e + 2d\sigma/f)}$$

where $c^2 = \frac{1}{2}\left[(\varepsilon - \sin^2\theta) + \sqrt{\{(\varepsilon - \sin^2\theta)^2 + 4\sigma^2/f^2\}}\right]$

$$d^2 = \frac{1}{2}\left[(-\varepsilon - \sin^2\theta) + \sqrt{\{(\varepsilon - \sin^2\theta)^2 - 4\sigma^2/f^2\}}\right]$$

When $\varepsilon = 0$

$$K_i = \frac{2(\sigma/f)\cos^2\theta - 1}{2(\sigma/f)\cos^2\theta + 2\sqrt{\{(\sigma/f\cos\theta) + 1\}}}$$

(iv) Vertical coil. Vector parallel to X axis:—

$$f_\theta + jf'_\theta = \frac{j2\pi}{\lambda} \sin \phi \delta A; \quad f_\phi + jf'_\phi = -\frac{j2\pi}{\lambda} \cos\theta \cos\phi \delta A \quad (10)$$

(v) Vertical coil. Vector parallel to Y axis:—

$$f_\theta + jf'_\theta = -j\frac{2\pi}{\lambda} \cos \phi \delta A;$$

$$f_\phi + jf'_\phi = -j\frac{2\pi}{\lambda} \cos\theta \sin \phi \delta A \quad . \quad (11)$$

(c) *Special Cases.*

The above formulæ are seldom required in their complete form. The following are the more usual:

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(vi) Horizontal coil. Vector parallel to Z axis:—

$$f_\theta + jf'_\theta = 0; \quad f_\phi + jf'_\phi = j\frac{2\pi}{\lambda}\sin\theta\delta A \quad . \quad (12)$$

A vertical current element can only produce vertically polarized components, but a horizontal current element may produce either and usually produces a combination of both. This is readily seen from equations (7), (8) and (9). A horizontal element will only produce purely horizontally polarized waves in a plane perpendicular to the plane in which the element lies, while it will produce purely vertically polarized waves in the plane in which it lies. In any other plane it will produce a combination of horizontally and vertically polarized waves.

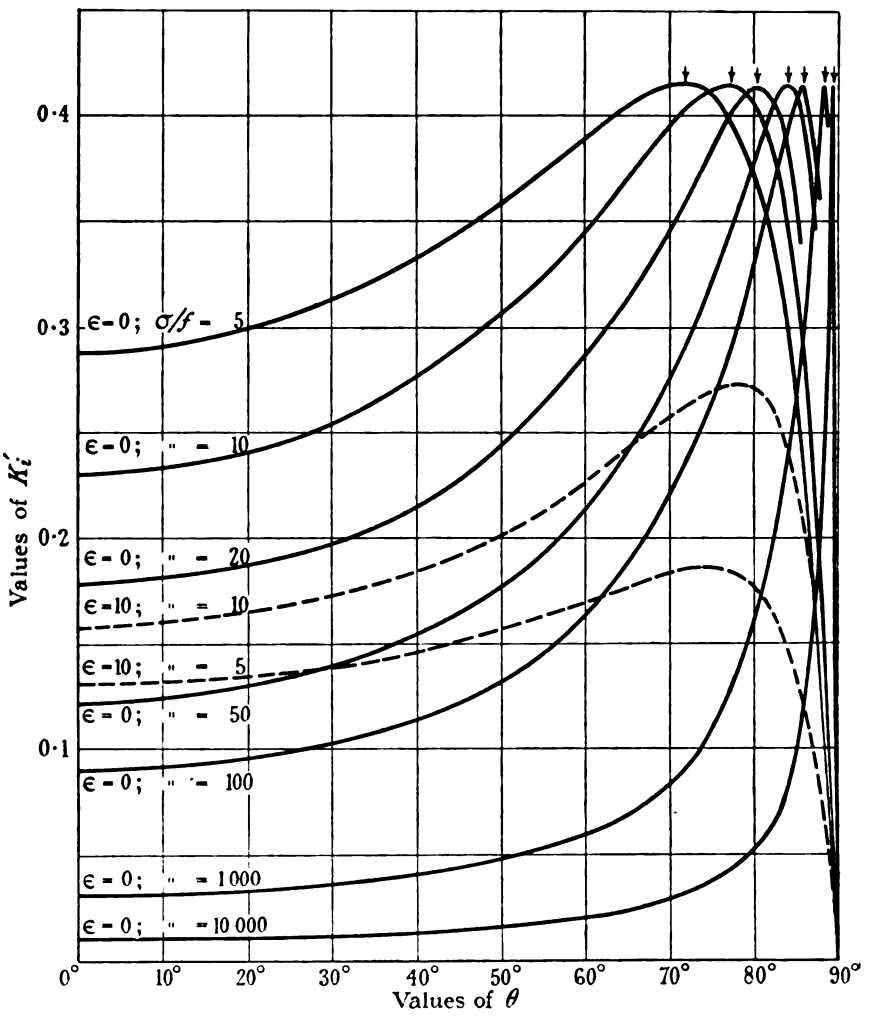
0·4
0·3
0·2
0·1
Values of K'_i
$\epsilon = 0$; $\sigma/f = 5$
$\epsilon = 0$; " = 10
$\epsilon = 0$; " = 20
$\epsilon = 10$; " = 10
$\epsilon = 10$; " = 5
$\epsilon = 0$; " = 50
$\epsilon = 0$; " = 100
$\epsilon = 0$; " = 1000
$\epsilon = 0$; " = 10 000
0° 10° 20° 30° 40° 50° 60° 70° 80° 90°
Values of θ

FIG. 5.

$$K_i = \frac{2\cos\theta(2c\sigma/f - \varepsilon l)}{(\varepsilon^2 + 4\sigma^2/f^2)\cos^2\theta + (c^2 + d^2) + 2\cos\theta(\varepsilon c + 2d\sigma/f)}$$

where $c^2 = \frac{1}{2}\left[(\varepsilon\sin^2\theta) + \sqrt{\{(\varepsilon - \sin^2\theta)^2 + 4\sigma^2/f^2\}}\right]$

$$d^2 = \frac{1}{2}\left[-(\varepsilon - \sin^2\theta) + \sqrt{\{(\varepsilon - \sin^2\theta)^2 + 4\sigma^2/f^2\}}\right]$$

When $\varepsilon = 0$

$$K'_i = \frac{2\sqrt{(\sigma/f)}\cos\theta}{2(\sigma/f)\cos^2\theta + 2\sqrt{(\sigma/f)}\cos\theta + 1}$$

giving maximum of $K' = 0{\cdot}414$ when $\cos\theta = \sqrt{(f/2\sigma)}$

(3) POLARIZATION.

The propagation of electromagnetic waves in or on the surface of an absorbing medium depends considerably on the polarization of the waves. In equations (1) and (2) E_θ is usually known as the vertically polarized component and E_ϕ as the horizontally polarized component.

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(4) Current Distribution.

(a) Case of a Uniform Wire.

On the assumption of uniformly distributed inductance and capacity and negligible resistance it can readily be shown that the current distribution along a simple straight-wire antenna should be sinusoidal. That is, if the origin is taken to be the open end of the antenna, the current at a distance x from it will be

$$I \sin\left(\frac{2\pi x}{\lambda}\right) \quad . \quad . \quad . \quad . \quad . \quad (13)$$

This simple sinusoidal law of distribution of current was proved experimentally by the author * to hold with sufficient accuracy, except for local end-effects, for lengths up to $\frac{3}{4}\lambda$, and there is no reason to suppose that it will fail, except possibly when the total is very many wavelengths long.

In applying this formula it is necessary to be particularly careful, when there are discontinuities in the antenna such as multiple branches, a change in the diameter of the wire, lumped inductances, etc. On the simple theory, the voltage and current at any point distant x from the end of a length of uniform wire can be written in the form

$$v = jIa \cos\frac{\pi}{\lambda}(x + b) \quad . \quad . \quad . \quad (14)$$

and

$$i = I \sin\frac{2\pi}{\lambda}(x + b) \quad . \quad . \quad . \quad . \quad (15)$$

In these equations I depends on the amplitude of the current, while b depends on the type of impedance connected at the end of the wire from which the distance x is measured. (This impedance is generally of the form of one or several lengths of wire.) a is a constant depending on the distributed constants of the wire and is equal to $\sqrt{(L/C)}$, where L is the self-inductance per unit length and C the capacity per unit length. We may write

$$a = \sqrt{(L/C)} = \frac{1}{cC} = cL \quad . \quad . \quad . \quad (16)$$

since c, the velocity of light, is very nearly equal to $1/\sqrt{(LC)}$.

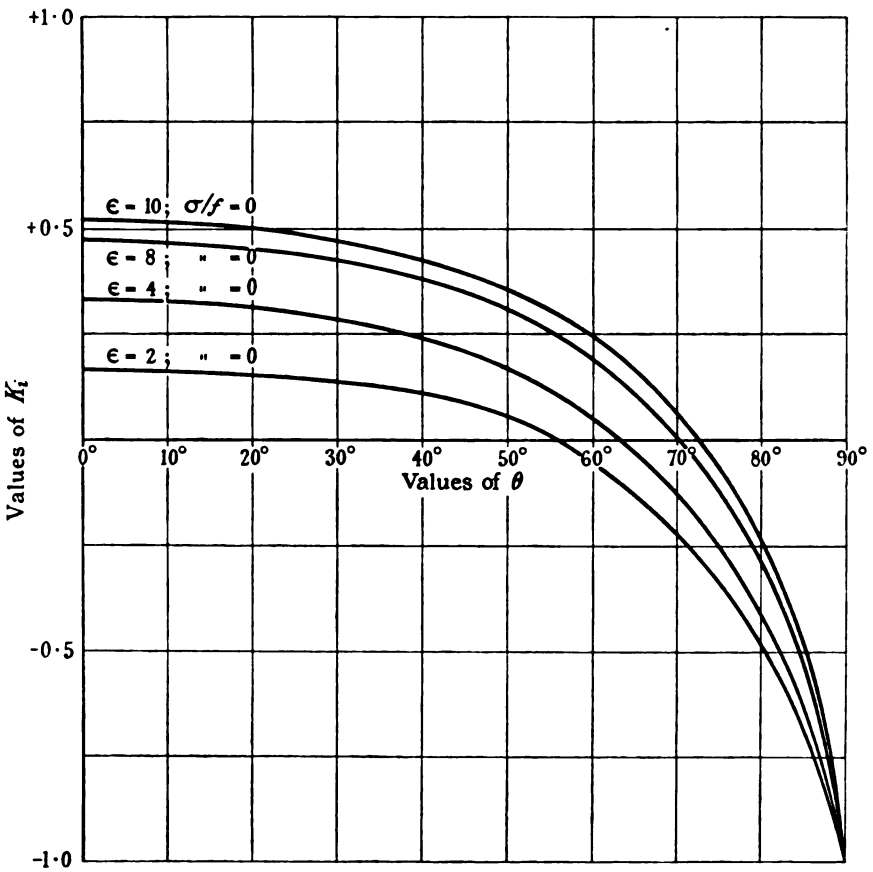

Fig. 6.

When $\sigma/f = 0$

$$K_i = \frac{\varepsilon \cos\theta - \sqrt{(\varepsilon - \sin^2\theta)}}{\varepsilon \cos\theta + \sqrt{(\varepsilon - \sin^2\theta)}}$$

$$K'_i = 0$$

* See Bibliography, (13).

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It will be noted that b has the dimensions of a length, so that the impedance at the end of the wire may be the point of discontinuity, together with a change in amplitude of the current.

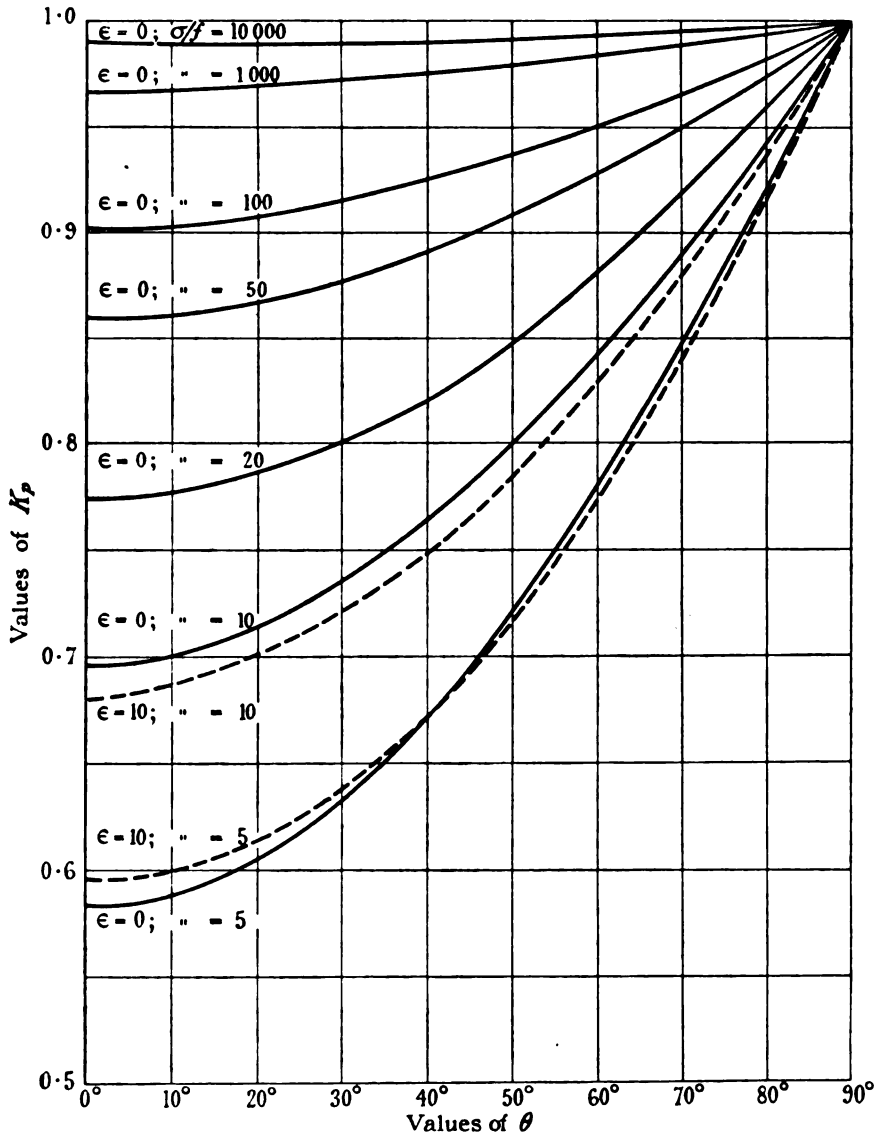

FIG. 7.

$$K_p = \frac{c^2 + d^2 - \cos^2\theta}{c^2 + d^2 + \cos^2\theta + 2c\cos\theta}$$

where $c^2 = \frac{1}{2}\left[(\varepsilon - \sin^2\theta) + \sqrt{\{(\varepsilon - \sin^2\theta^2) + 4\sigma^2/f^2\}}\right]$

$$d^2 = \frac{1}{2}\left[-(\varepsilon - \sin^2\theta) + \sqrt{\{(\varepsilon - \sin^2\theta) + 4\sigma^2/f^2\}}\right]$$

When $\varepsilon = 0$

$$K_p = \frac{2(\sigma/f) - \cos^2\theta}{2(\sigma/f) + \cos^2\theta + 2(\sigma/f)\cos\theta}$$

represented by a length so far as the current and potential along the wire are concerned. When there is no impedance at the end of the wire, that is when that end is open, $b = 0$. It will be shown that any discontinuity can be represented, as far as the current distribution is concerned, by a loss or a gain of length of the wire at

(*b*) *Case of n Branches meeting in a Point.*

Consider an antenna consisting of a single wire from the end of which n wires branch off. It does not matter what wires are connected to the end of these. Let the lengths of these branches be l_1, l_2 . . . and the equivalent lengths of the impedance at their ends be b_1, b_2 . . .

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Let the current and potential along the single wire, with the point O as origin, be respectively

$$I \sin \frac{2\pi}{\lambda}(x + b) \text{ and } jIa \cos \frac{2\pi}{\lambda}(x + b) \quad . \quad (17)$$

Then, if v_0 is the potential at the point O, we have

$$v_0 = jIa \cos\left(\frac{2\pi b}{\lambda}\right) = jI_1a_1 \cos \frac{2\pi}{\lambda}(l_1 + b_1)$$
$$= jI_2a_2 \cos \frac{2\pi}{\lambda}(l_2 + b_2) = \ldots \quad . \quad (18)$$

and

$$I \sin\left(\frac{2\pi b}{\lambda}\right) = I_1 \sin \frac{2\pi}{\lambda}(l_1 + b_1)$$
$$+ I_2 \sin \frac{2\pi}{\lambda}(l_2 + b_2) + \ldots \quad . \quad (19)$$

From equations (18) the ratios of the currents in the various branches can be calculated. Also, combining equations (17) and (18), we have

$$\frac{1}{a} \tan\left(\frac{2\pi b}{\lambda}\right) = \frac{1}{a_1} \tan \frac{2\pi}{\lambda}(l_1 + b_1)$$
$$+ \frac{1}{a_2} \tan \frac{2\pi}{\lambda}(l_2 + b_2) + \ldots \quad . \quad (20)$$

from which b can be readily obtained, thus giving the current distribution from equations (17).

The branches are commonly the same in diameter and of equal lengths l, in which case equations (19) and (20) reduce to

$$\left.\begin{aligned} I &= I_1 n \sin\left(\frac{2\pi l}{\lambda}\right) \sin\left(\frac{2\pi b}{\lambda}\right) \\ \text{and} \quad \tan\left(\frac{2\pi b}{\lambda}\right) &= n \tan\left(\frac{2\pi l}{\lambda}\right) \end{aligned}\right\} \quad . \quad (21)$$

In the special case when $\tan(2\pi l/\lambda) = 0$, $I = I_1$.

Thus, at the point of discontinuity there is an effective loss of length equal to $(l - b)$ as far as the current distribution is concerned.

(c) *Effect of a Lumped Inductance.*

Suppose that at a distance l along the length of an antenna there is a self-inductance L_0 having negligible capacity. Let the current and potential above this inductance be given by

$$i = I \sin \frac{2\pi}{\lambda}(l + b) \quad \text{and} \quad v = jIa \cos \frac{2\pi}{\lambda}(l + b)$$

while that below the inductance is given by

$$i' = I' \sin \frac{2\pi}{\lambda}(l + b') \quad \text{and} \quad v' = jI'a' \cos \frac{2\pi}{\lambda}(l + b')$$

Since $i = i'$ and $v - v' = jL_0\omega i$, we have

$$\left.\begin{aligned} \cot \frac{2\pi}{\lambda}(l + b') &= \frac{L}{L'} \cot \frac{2\pi}{\lambda}(l + b) - \frac{2\pi}{\lambda} \cdot \frac{L_0}{L'} \\ &= \frac{C'}{C} \cot \frac{2\pi}{\lambda}(l + b) - \frac{2\pi}{\lambda} c^2 C' L_0 \end{aligned}\right\} \quad (22)$$

while

$$\frac{I'}{I} = \frac{\sin 2\pi(l + b')}{\sin 2\pi(l + b)} \quad . \quad . \quad . \quad . \quad (23)$$

where L, C and L', C' are the distributed constants of the wire above and below the self-inductance respectively. Thus the effect of the discontinuity is to change the ratio of the maximum currents in the ratio given by equation (23) and at the same time produce the effect as though a length of antenna wire equal to $(b' - b)$ were concentrated at the point of discontinuity as given by equations (22).

In most cases the wire will have the same diameter above and below the lumped inductance, in which case $L = L'$ and $C = C'$.

(d) *Effect of a Change in the Distributed Constants.*

If there be a sudden change in the diameter of the wire, the effect of the discontinuity can be obtained directly by putting $L_0 = 0$ in equations (22), whence we obtain

$$\cot \frac{2\pi}{\lambda}(l + b') = \frac{L}{L'} \cot \frac{2\pi}{\lambda}(l + b) = \frac{C'}{C} \cot \frac{2\pi}{\lambda}(l + b) \quad (24)$$

and

$$\frac{I'}{I} = \frac{\sin 2\pi(l + b')}{\sin 2\pi(l + b)} \quad . \quad . \quad . \quad . \quad (25)$$

It is difficult to calculate accurately the value of C/C' or L/L', chiefly because the exact meaning of the distributed constants is somewhat vague and their use is essentially dependent on an approximation.* We may obtain a sufficiently good approximation by calculating the antenna capacity in two cases, assuming first that the distributed constants are L and C over the whole antenna and secondly that the distributed constants are L' and C'. It will generally be a sufficiently close approximation for this purpose to use the simple relation

$$\frac{L}{L'} = \frac{C'}{C} = \frac{\log(2l/d)}{\log(2l/d')} \quad . \quad . \quad . \quad . \quad (26)$$

where d and d' are the diameters of the wire and l the total length of the antenna. If the antenna is several wavelengths long, it will be close enough to assume $l = \frac{1}{4}\lambda$.

It will be seen that, if the diameter of the wire is greater *above* than it is *below* the point of discontinuity, there is an effective *loss* of length at that point, and vice versa.

(5) Effect of Image in the Ground.

The full formulæ for calculating the effect of an imperfect earth have been given by the author.† Unfortunately, owing to an unnoticed ambiguity, errors were made in some of the signs. The corrected equations are given below. On this theory, which was first suggested by T. L. Eckersley ‡ and amplified by the author,§ an element of current $I\delta l$ has for image a similar element as shown (Fig. 3) but of strength $(K - jK')I\delta l$, where K and K' are real functions of the angle of transmission or reception to the vertical and also depend on whether the element of current lies

* See Bibliography, (14). † *Ibid.*, (15). ‡ *Ibid.*, (16). § *Ibid.*, (17).

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in the plane or perpendicular to that in which propagation is considered, i.e. on the polarization of the ray.

In the special case that is approached at long waves and over water when the earth acts as a perfect conductor, $(K - jK') = 1$.

When the element of current lies in the plane of propagation under consideration K and K' are represented by K_i and K_i' and have values given by

$$K_i - jK_i' = \frac{(\epsilon^2 + 4\sigma^2/f^2)\cos^2\theta - (c^2 + d^2) - 2j\cos\theta\{2c(\sigma/f) - \epsilon d\}}{(\epsilon^2 + 4\sigma^2/f^2)\cos^2\theta + c^2 + d^2 + 2\cos\theta(\epsilon c + 2d\sigma/f)} \quad . \quad . \quad . \quad . \quad . \quad (27)$$

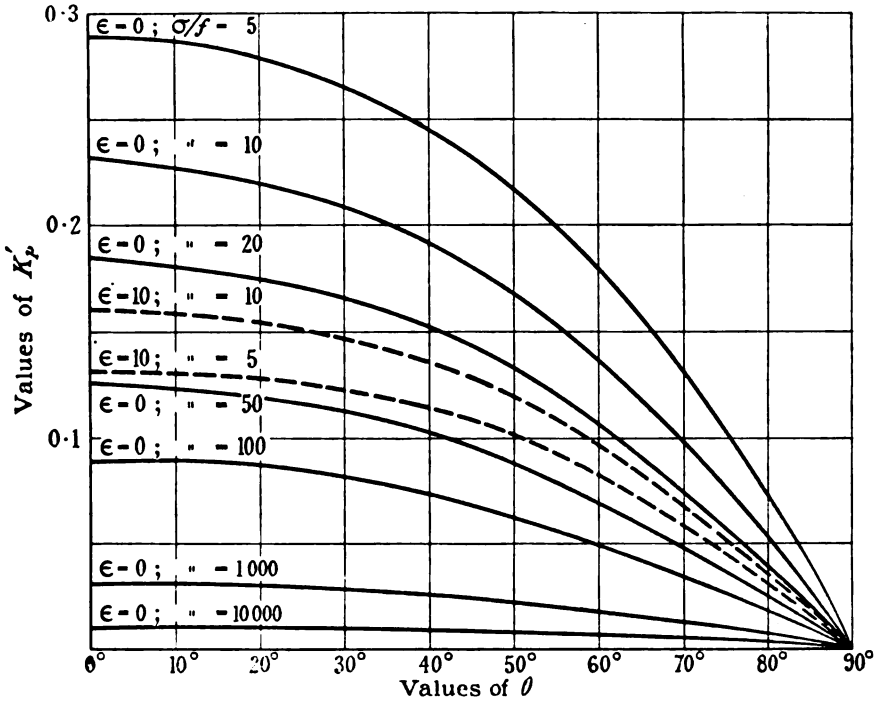
0·3
0·2
0·1
Values of K_p'
$\epsilon = 0$; $\sigma/f = 5$
$\epsilon = 0$; " = 10
$\epsilon = 0$; " = 20
$\epsilon = 10$; " = 10
$\epsilon = 10$; " = 5
$\epsilon = 0$; " = 50
$\epsilon = 0$; " = 100
$\epsilon = 0$; " = 1000
$\epsilon = 0$; " = 10000
0° 10° 20° 30° 40° 50° 60° 70° 80° 90°
Values of θ

FIG. 8.

$$K_p' = \frac{2d\cos\theta}{c^2 + d^2 + \cos^2\theta + 2c\cos\theta}$$

where $c^2 = \frac{1}{2}\left[(\epsilon - \sin^2\theta) + \sqrt{\{(\epsilon - \sin^2\theta)^2 + 4\sigma^2/f^2\}}\right]$

$$d^2 = \frac{1}{2}\left[-(\epsilon - \sin^2\theta) + \sqrt{\{(\epsilon - \sin^2\theta)^2 + 4\sigma/f^2\}}\right]$$

When $\epsilon = 0$

$$K_p' = \frac{2(\sigma/f)\cos\theta}{2(\sigma/f) + \cos^2\theta + 2(\sigma/f)\cos\theta}$$

where σ is the conductivity in electrostatic units, ϵ is the dielectric constant, f is the frequency, $j = \sqrt{-1}$, and c and d are given by $c^2 - d^2 = \epsilon - \sin^2\theta$ and $cd = \sigma/f$, c and d being both real and positive. These two equations are easily solved by using Pierce's tabulated f and g functions.*

When σ/f can be neglected in comparison with ϵ this expression reduces to

$$K_i - jK_i' = \frac{\epsilon\cos\theta - \sqrt{(\epsilon - \sin^2\theta)}}{\epsilon\cos\theta + \sqrt{(\epsilon - \sin^2\theta)}} \quad . \quad (28)$$

* See Bibliography, (18).

When ϵ and $\sin^2\theta$ can be neglected in comparison with σ/f, the expression reduces to

$$K_i - jK_i' = \frac{(2\sigma/f)\cos^2\theta - 1 - 2j(\sigma/f)^{\frac{1}{2}}\cos\theta}{(2\sigma/f)\cos^2\theta + 1 + 2(\sigma/f)^{\frac{1}{2}}\cos\theta} \quad (29)$$

In the same way, when the plane of propagation under consideration is perpendicular to the current element, $K - jK'$ is written $K_p - jK_p'$ and is given by

$$K_p - jK_p' = \frac{c^2 + d^2 - \cos^2\theta - 2jd\cos\theta}{c^2 + d^2 + \cos^2\theta + 2c\cos\theta} \quad (30)$$

When σ/f can be neglected in comparison with ϵ this expression becomes

$$K_p - jK_p = \frac{\sqrt{(\epsilon - \sin^2\theta)} - \cos\theta}{\sqrt{(\epsilon - \sin^2\theta)} + \cos\theta} \quad . \quad (31)$$

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When ϵ and $\sin^2\theta$ can be neglected in comparison with σ/f the expression reduces to

$$K_p - jK_p' = \frac{(2\sigma/f) - \cos^2\theta - 2j(\sigma/f)^{\frac{1}{2}}\cos\theta}{(2\sigma/f) + \cos^2\theta + 2(\sigma/f)^{\frac{1}{2}}\cos\theta} \quad (32)$$

These expressions are somewhat laborious to evaluate, and a number of curves have been calculated. These are shown in Figs. 4 to 9. It may be noted that the value of K_i is not very much affected by the dielectric constant of the ground unless this is larger than $2\sigma/f$, but this is not so for K_i'. This means that the amplitude of the reflected ray will not in general be very much affected by the dielectric constant but that the phase of this ray will be affected to a much greater extent. This leads incidentally to the deduction that, if a method of measuring the dielectric constant of the ground based on the electromagnetic wave propagation is developed, it is desirable that it should depend on a phase measurement rather than on amplitude, unless the measurement is made in the neighbourhood of Brewster's angle.

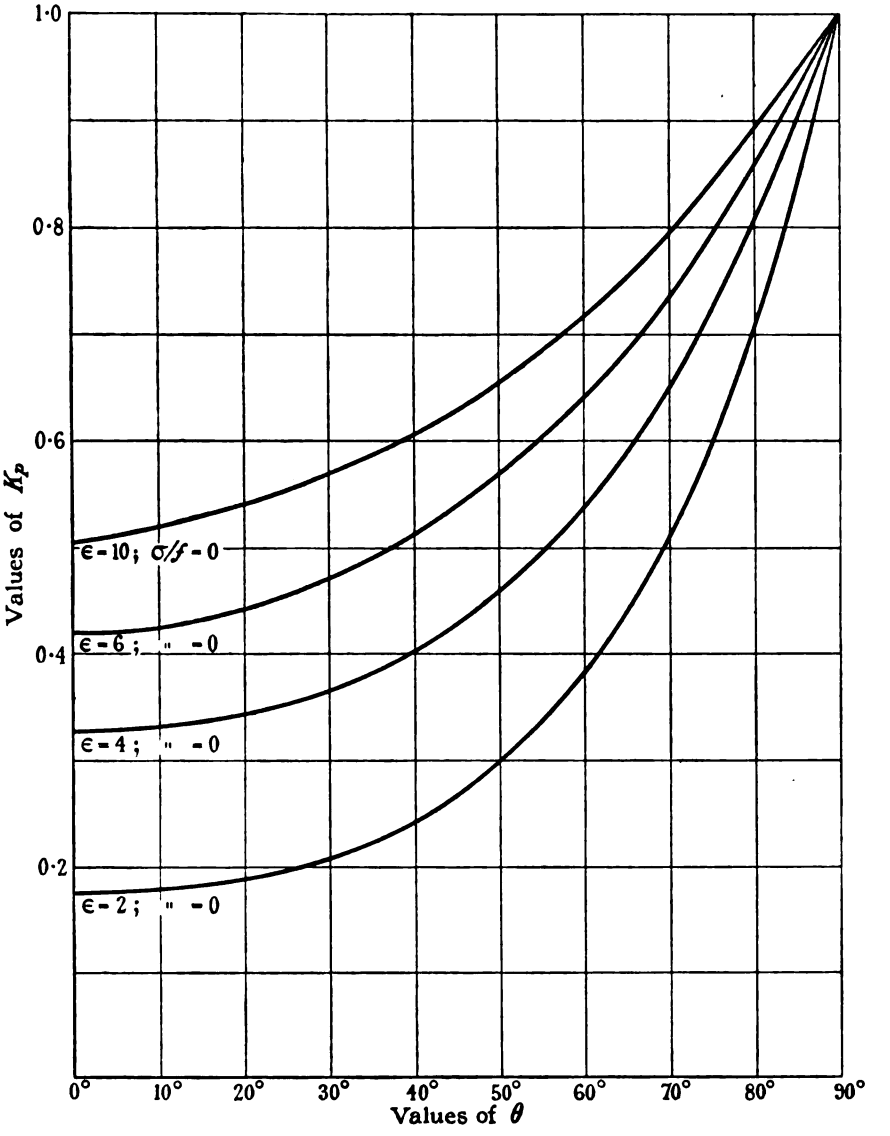

FIG. 9.

When $\sigma/f = 0$

$$K_p = \frac{\sqrt{(\varepsilon - \sin^2\theta)} - \cos\theta}{\sqrt{(\varepsilon - \sin^2\theta)} + \cos\theta}$$

$$K_p' = 0$$

From the curves given in Figs. 4 to 9 we can readily deduce the field from any element of current over an imperfect earth. Consider the field at a distant point P the radius vector of which makes an angle θ with the

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vertical. If h is the height of a vertical element above the ground we have, for the field E_v at P,

$$E_v = \frac{60\pi I \delta l \sin\theta}{\lambda r} \left\{(1 + K - jK')\cos\gamma + j(1 - K + jK')\sin\gamma\right\} \quad (33)$$

where $\gamma = \frac{2\pi h \cos\theta}{\lambda}$

Therefore

$$|E_v| = \frac{60\pi I \delta l \sin\theta}{\lambda r} \sqrt{[1 + K^2 + K'^2 + 2K\cos^2\gamma - 2K'\sin 2\gamma]} \quad (34)$$

In the case of a horizontal element lying in a plane perpendicular to the direction of transmission or reception it can be readily shown that the field E_h is given by

$$|E_h| = \frac{60\pi I \delta l \cos\theta}{\lambda r} \sqrt{[1 + K^2 + K'^2 - 2K\cos 2\gamma + 2K'\sin 2\gamma]} \quad (35)$$

By differentiation it may be shown that E_v is a maximum and E_h is a minimum when

$$2\gamma = (2n - 1)\pi - \text{arc tan } (K'/K) \quad (36)$$

while E_v is a minimum and E_h a maximum when

$$2\gamma = 2n\pi - \text{arc tan } (K'/K) \quad (37)$$

where n is an integer. It should be noted that K may either be K_i or K_p, depending on the direction of the antenna element, and similarly for K'.

As K'/K is always positive except for angles with the vertical greater than Brewster's angle, we should expect all those maxima and minima of the radiation distribution, which are produced by the interference of the antenna with its image, to tend to shift towards the Brewster's angle as the conductivity of the ground decreases. In any given case the amount of this shift should be a measure of the ratio K'/K, from which the conductivity of the ground might be deduced.

(6) Radiation Distribution of a Straight Wire Inclined at Any Angle.

(a) *General Case.*

For the computation of many of the formulæ the curves given in Figs. 12 and 13 will be found useful.

Consider the straight wire AB (Fig. 10) lying in the plane $\phi = \beta$. Let this antenna be of length l and let it make an angle α with the vertical, its lowest point A having co-ordinates $(r_0\cos\beta, r_0\sin\beta, z_0)$. That is, its lowest point is at a height z_0 and its projection on to the horizontal plane is at a distance r_0 from the origin. To make the problem as general as possible, we shall assume that the antenna is loaded at its end B with a load equivalent to a length l_1 of the antenna, so that the current, i, flowing in the direction AB, at any point Q on the antenna, distant z from A, is given by

$$i = I_0 \sin\frac{2\pi}{\lambda}(l + l_1 - z)\sin\omega t \quad (38)$$

The general formula, which will be given below, can be used for any individual straight portion of an antenna, and, by adding the field due to all these portions which form the whole antenna system, it is possible to obtain an expression for the radiation distribution of any antenna system. In complicated systems, however, the expressions are much too cumbersome and a combination of formulæ and graphical methods becomes more suitable.

Consider the distant point P having its horizontal projection at N. Let PO make an angle ψ with the antenna AB. If the distance of the point P is large compared with the length of the antenna, the value of ψ will not vary materially, whatever be the position of the point Q on AB, and we may assume that the angle ψ is constant, so that

$$\cos\psi = \sin\alpha\sin\theta\cos(\phi - \beta) + \cos\alpha\cos\theta \quad (39)$$

Now the field at P due to the element of wire of length δz at Q is inversely proportional to the distance PQ. As before, we may assume that the distance PQ is independent of the position of Q on AB and is equal to r. We must, however, take into account the phase at which the radiation from Q reaches P, for this is dependent on the position of Q, unless the length AB is small compared with the wavelength. A current at the origin in phase with that at Q would send a radiation which would arrive at P, τ seconds after that from Q, where

$$c\tau = \text{PO} - \text{PQ} = d + z\cos\psi \quad (40)$$

where d is the length of the projection of OA on OP and is given by

$$d = x_0\sin\theta\cos(\phi - \beta) + z_0\cos\theta \quad (41)$$

From equations (3), (4) and (38) we obtain for the radiation distribution of the whole antenna AB that

$$f_\theta + jf'_\theta = \int_0^l \sin\frac{2\pi}{\lambda}(l + l_1 - z)\left\{\cos\alpha\sin\theta - \sin\alpha\cos\theta\cos(\phi - \beta)\right\}\left\{\cos\pi l(d + z\cos\psi) + j\sin\frac{2\pi}{\lambda}(d + z\cos\psi)\right\}dz$$

and

$$f_\phi + jf'_\phi = \int_0^l \sin\frac{2\pi}{\lambda}(l + l_1 - z)\sin\alpha\sin(\phi - \beta)\left\{\cos\frac{2\pi}{\lambda}(d + z\cos\psi) + j\sin\frac{2\pi}{\lambda}(d + z\cos\psi)\right\}dz$$

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where the current is measured at the mathematical current maximum, even if such a maximum does not occur within the length AB.

If F is the resultant of $(f_\theta + jf'_\theta)$ and $(f_\phi + jf'_\phi)$, so that

$$F^2 = (f_\theta + jf'_\theta)^2 + (f_\phi + jf'_\phi)^2$$

then the above integrals reduce to

$$F = \frac{\lambda}{2\pi \sin\psi}\Big[\cos A \cos D - \cos B \cos C - \cos\psi\,(\sin A \sin D - \sin B \sin C) + j\{\cos A \sin D - \cos B \sin C + \cos\psi\,(\sin A \cos D - \sin B \cos C)\}\Big] \quad . \quad (42)$$

$$f_\theta + jf'_\theta = \frac{\{\cos\alpha \sin\theta - \sin\alpha \cos\theta \cos(\phi - \beta)\}}{\sin\psi} F \quad (43)$$

and

$$f_\phi + jf'_\phi = -\frac{\sin\alpha \sin(\phi - \beta)}{\sin\psi} F \quad . \quad . \quad (44)$$

where $A = 2\pi l_1/\lambda$
$B = 2\pi(l + l_1)/\lambda$
$C = 2\pi d/\lambda$
$D = 2\pi(d + l\cos\psi)/\lambda$

The values of $\cos\psi$ and d are given by equations (39) and (41). The sign of $\cos\psi$ will come automatically from equation (39), while that of $\sin\psi$ is immaterial so long as it is taken consistently the same, for $\sin\psi$ only occurs in the final expression for the field strength as a square.

In order to find the effect of the image it is necessary to put $(\pi - \alpha)$ for α, $-z_0$ for z_0 in the above expressions, and multiply the result by the factor $-(K - jK')$ as explained in section (5) above and given in Figs. 4 to 9. The result should then be added to the radiation distribution for the antenna without its image in the ground.

These expressions as they stand are much too complex for practical use. They reduce considerably, however, in a large number of special cases, some of which will now be considered.

(b) Vertical Antenna Loaded at the Top.

In this case $\alpha = 0$ and $x_0 = 0$, so that $d = z_0 \cos\theta$ and $\cos\psi = \cos\theta$.

Hence

$$f_\theta + jf'_\theta = \frac{\lambda}{2\pi \sin\theta}\Bigg[\cos\frac{2\pi l_1}{\lambda}\left\{\cos\frac{2\pi\cos\theta(l + z_0)}{\lambda} + j\sin\frac{2\pi\cos\theta(l + z_0)}{\lambda}\right\} - \cos\frac{2\pi(l + l_1)}{\lambda}\left\{\cos\frac{2\pi z_0\cos\theta}{\lambda} + j\sin\frac{2\pi z_0\cos\theta}{\lambda}\right\} - \cos\theta\left[\sin\frac{2\pi l_1}{\lambda}\left\{\sin\frac{2\pi\cos\theta(l + z_0)}{\lambda} - j\cos\frac{2\pi\cos\theta(l + z_0)}{\lambda}\right\} - \sin\frac{2\pi(l + l_1)}{\lambda}\left\{\sin\frac{2\pi z_0\cos\theta}{\lambda} - j\cos\frac{2\pi z_0\cos\theta}{\lambda}\right\}\right]\Bigg]^* \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (45)$$

and $f_\phi + jf'_\phi = 0$.

The expression for the image, if we assume a perfect earth, is identical with expression (45), except that $(l + z_0)$ becomes $(l - z_0)$ and all the imaginary terms have their signs reversed, so that, if $z_0 = 0$, the total effect of the antenna and its image together is to double the real part of equation (45) and bring the imaginary part to zero.

If there is no load at the top and the antenna reaches to the ground (i.e. $l_1 = 0$ and $z_0 = 0$), expression (45) reduces (not taking the image into account) to

$$f_\theta + jf'_\theta = \frac{\lambda}{2\pi \sin\theta}\left[\cos\frac{2\pi l\cos\theta}{\lambda} - \cos\frac{2\pi l}{\lambda} + j\left\{\sin\frac{2\pi l\cos\theta}{\lambda} - \cos\theta\sin\frac{2\pi l}{\lambda}\right\}\right] \quad . \quad (46)$$

(c) Horizontal Antenna not Loaded at the End.

In this case $\alpha = \frac{1}{2}\pi$, $x_0 = 0$, $l_1 = 0$.

Hence $F = \frac{\lambda}{2\pi \sin\psi}\Bigg[\cos\frac{2\pi(z_0\cos\theta + l\cos\psi)}{\lambda} - \cos\frac{2\pi l}{\lambda}\cos\frac{2\pi z_0\cos\theta}{\lambda} + \cos\psi\sin\frac{2\pi l}{\lambda}\sin\frac{2\pi z_0\cos\theta}{\lambda} + j\left\{\sin\frac{2\pi(z_0\cos\theta + l\cos\psi)}{\lambda} - \cos\frac{2\pi l}{\lambda}\sin\frac{2\pi z_0\cos\theta}{\lambda} - \cos\psi\sin\frac{2\pi l}{\lambda}\cos\frac{2\pi z_0\cos\theta}{\lambda}\right\}\Bigg] \quad . \quad . \quad . \quad . \quad (47)$

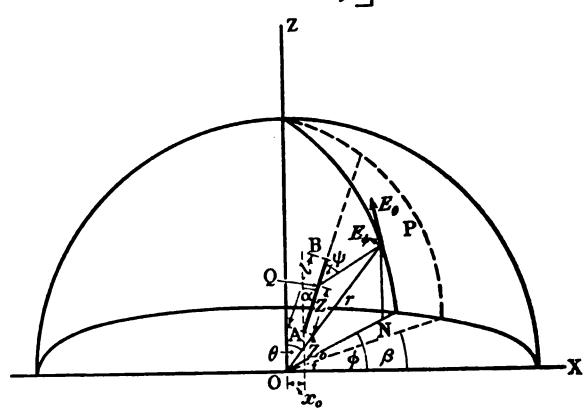

FIG. 10.

* Curves of $\cos\{2\pi(l/\lambda)\cos\theta\}$, $\cos\{2\pi(l/\lambda)\sin\theta\}$, $\sin\{2\pi(l/\lambda)\sin\theta\}$ and $\sin\{2\pi(l/\lambda)\cos\theta\}$ for various values of l/λ are given in Figs. 12 and 13.

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$$f_\theta + jf'_\theta = -\frac{\cos\theta\cos(\phi-\beta)}{\sin\psi}F$$

and $$f_\phi + jf'_\phi = -\frac{\sin(\phi-\beta)}{\sin\psi}F \quad \Big\} \quad . \quad . \quad (48)$$

where $\cos\psi = \sin\theta\cos(\phi-\beta)$.

The expression for the image, if we assume a perfect earth, is identical with expression (47), except that z_0 becomes $-z_0$ and F becomes $-F$. The expression for the total field due to the antenna and its image in the earth will be given by equations (48), where

$$F = \frac{\lambda}{\pi\sin\psi}\sin\frac{2\pi z_0\cos\theta}{\lambda}\left[\cos\psi\sin\frac{2\pi l}{\lambda} - \sin\frac{2\pi l\cos\psi}{\lambda} + j\left\{\cos\frac{2\pi l\cos\psi}{\lambda} - \cos\frac{2\pi l}{\lambda}\right\}\right] \quad (49)$$

By adding the expressions in this section to the suitable expressions from section (6) (*b*), the radiation distribution of any inverted "L" antenna can be obtained.

(*d*) *Two Equal Horizontal Antennæ in One Straight Line with Currents Flowing in Opposite Directions.*

If we put $(\pi+\beta)$ for β in equations (47), (48) and (49) we shall obtain the expression for an equal horizontal antenna with current flowing in the opposite direction to that in the first. The two expressions can be added together, giving the formula required.

For the antenna alone we have, together with equations (48),

$$F = \frac{\lambda}{\pi\sin\psi}\left\{\cos\psi\sin\frac{2\pi l}{\lambda} - \sin\frac{2\pi l\cos\psi}{\lambda}\right\}\left\{\sin\frac{2\pi z_0\cos\theta}{\lambda} - j\cos\frac{2\pi z_0\cos\theta}{\lambda}\right\} \quad . \quad (50)$$

The expression for the total antenna with its image in a perfect earth is given by equation (48), where

$$F = -\frac{2\lambda}{\pi\sin\psi}\sin\left(\frac{2\pi z_0\cos\theta}{\lambda}\right)\left\{\cos\psi\sin\left(\frac{2\pi l}{\lambda}\right) - \sin\left(\frac{2\pi l\cos\psi}{\lambda}\right)\right\} \quad . \quad (51)$$

It should be noted that the current measured is the mathematical maximum of either limb of the antenna. By combining the expressions in this section with the suitable expressions from section (6)(*b*), the radiation distribution of any "T" antenna can be obtained. In doing this the effect of discontinuity produced by the double branch at the top of the vertical portion must be taken into account. For this purpose equations (21) should be used.

(*e*) *Two Equal Horizontal Antennæ in One Straight Line with Current Flowing in the Same Direction.*

This is similar to the problem of the preceding paragraph. The required expression is obtained by putting $(\pi+\beta)$ for β in equations (47), (48) and (49) and subtracting the two expressions in each case.

For the antenna alone we have, in combination with equation (48),

$$F = \frac{\lambda}{\pi\sin\psi}\left\{\cos\frac{2\pi l\cos\psi}{\lambda} - \cos\frac{2\pi l}{\lambda}\right\}\left\{\cos\frac{2\pi z_0\cos\theta}{\lambda} + j\sin\frac{2\pi z_0\cos\theta}{\lambda}\right\} \quad . \quad (52)$$

The expression for the total antenna and its image in a perfect earth is given by equations (48), where

$$F = j\frac{2\lambda}{\pi\sin\psi}\sin\frac{2\pi z_0\cos\theta}{\lambda}\left\{\cos\frac{2\pi l\cos\psi}{\lambda} - \cos\frac{2\pi l}{\lambda}\right\} \quad . \quad (53)$$

These expressions represent the radiation distribution of a horizontal antenna excited at the centre.

(*f*) *General Case when Current is Uniform instead of Sinusoidal.*

It has been pointed out by the author * that an increase in directivity could be obtained by reducing the space taken by those portions of the antenna in which the current is small. This can be done in practice by means of the device employed by Franklin in which the antenna wire is bent on itself several times, so that large portions of the current distribution for a simple straight wire are concentrated into a relatively small space. Carrying this out to the extreme would result in a nearly uniform current distribution along the antenna instead of the usual sinusoidal distribution. It may be of interest for design purposes to find whether the attainment of a more uniform current distribution is of value. For this purpose it is necessary to know what is the limiting condition that can thus be obtained. Such a condition can, of course, never be actually reached, but in many cases it could be very nearly obtained. The following formulæ refer to such limiting cases, in which the current is assumed perfectly uniform over the whole antenna.

The general case gives the following:—

$$f_\theta + jf'_\theta = \frac{\cos\alpha\sin\theta - \sin\alpha\cos\theta\cos(\phi-\beta)}{\sin\psi}F \quad (54)$$

$$f_\phi + jf'_\phi = -\frac{\sin\alpha\sin(\phi-\beta)}{\sin\psi}F \quad . \quad . \quad (55)$$

where

$$F = \frac{\lambda}{2\pi}\tan\psi\left[\sin\frac{2\pi(d+l\cos\psi)}{\lambda} - \sin\frac{2\pi d}{\lambda} + j\left\{\cos\frac{2\pi d}{\lambda} - \cos\frac{2\pi(d+l\cos\psi)}{\lambda}\right\}\right] \quad . \quad (56)$$

$$\cos\psi = \sin\alpha\sin\theta\cos(\phi-\beta) + \cos\alpha\cos\theta$$

and $$d = x_0\sin\theta\cos(\phi-\beta) + z_0\cos\theta$$

As before, F is the resultant of $(f_\theta + jf'_\theta)$ and $(f_\phi + jf'_\phi)$.

* See Bibliography, (19).

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(7) A System of Similar Parallel Antennæ Spaced in any Given Manner.

(*a*) *General Case.*

It often occurs in a system of antennæ that there are a number of parallel portions of equal length and carrying similar distributions of current. Instead of adding the field due to each portion separately, we may proceed as follows.

Let there be n such portions and let O_1, O_2, . . ., O_n be points on the 1st, 2nd, . . ., nth portion respectively. These points are in the same relative position on each individual portion. Taking a single portion we can calculate its radiation distribution from the equations

$$F + jF' = \frac{(f + jf')}{n}\left[\frac{\cos\left\{\beta_1 + \frac{2\pi d_1}{\lambda} + (n-1)\left(\frac{\beta}{2} + \frac{\pi d}{\lambda}\right)\right\}\sin n\left(\frac{\beta}{2} + \frac{\pi d}{\lambda}\right)}{\sin\left(\frac{\beta}{2} + \frac{\pi d}{\lambda}\right)}\right.$$
$$\left. + j\frac{\sin\left\{\beta_1 + \frac{2\pi d_1}{\lambda} + (n-1)\left(\frac{\beta}{2} + \frac{\pi d}{\lambda}\right)\right\}\sin n\left(\frac{\beta}{2} + \frac{\pi d}{\lambda}\right)}{\sin\left(\frac{\beta}{2} + \frac{\pi d}{\lambda}\right)}\right] \quad . \quad (60)$$

given in section (6). Let this be $(f + jf')$, this expression representing in fact two functions, one for each component in space of the field.

Supposing that the co-ordinates of O_1, O_2, . . ., O_n are (x_1, y_1, z_1), (x_2, y_2, z_2), . . . (x_n, y_n, z_n) respectively, and that the currents I_1, I_2, . . . I_n flowing at these points are in advance on some arbitrary phase by β_1, β_2, . . ., β_n respectively, we may write for the radiation distribution $(F + jF')$ of the whole system

$$F + jF' = (f + jf')\left[\frac{\sum_{r=1}^{r=n} I_r \cos\left\{\beta_r + \frac{2\pi d_r}{\lambda}\right\} + j\sum_{r=1}^{r=n} I_r \sin\left\{\beta_r + \frac{2\pi d_r}{\lambda}\right\}}{\sum_{r=1}^{r=n} I_r}\right] \quad (57)$$

where $d_1, d_2, \ldots, d_n$ are the projections of OO_1, OO_2, . . ., OO_n on the direction considered, O being the origin. Thus, if the direction considered is (θ, ϕ),

$$d_r = (x_r \cos\phi + y_r \sin\phi)\sin\theta + z_r \cos\theta \quad . \quad (58)$$

If there are not several systems to add together, it is unnecessary to consider the phase of the radiation distribution. Hence, if f_0 is the resultant of $(f + jf')$ and F_0 that of $(F + jF')$, equation (57) reduces to

$$F_0^2 = f_0^2 \frac{\sum_{r=1}^{r=n}\sum_{s=r}^{s=n}\left[I_r I_s \cos\left\{\beta_r - \beta_s + \frac{2\pi}{\lambda}(d_r - d_s)\right\}\right]}{\left(\sum_{r=1}^{r=n} I_r\right)^2} \quad . \quad (59)$$

It was mentioned in section (1) that the point of the antenna system at which the current is measured should be stated; in this particular case the current is not measured at a single point, but it is taken as the sum (irrespective of phase) of all the currents at the points O_1, O_2, . . ., O_n.

(*b*) *Uniform Spacing, Uniform Variation of Phase and Equal Currents.*

This is a condition frequently met with in practice, and for this case equation (57) adopts a simpler form. If the portions of the antenna system are uniformly spaced, the projection of the lengths OO_1, OO_2, . . ., OO_n on to the direction required can be written in the form d_1, $d_1 + d$, . . ., $d_1 + (n-1)d$ respectively, the values of d_1 and d being readily obtained from equation (58). If, in the same way, the phases are β_1, $\beta_1 + \beta$, . . ., $\beta_1 + (n-1)\beta$ and the current is of the same amplitude in each portion of the antenna system, equation (57) becomes

If there are not several systems to add together, it is unnecessary to consider the phase of the radiation distribution. In this case the resultant, F_0, is given by

$$F_0 = \frac{f_0}{n} \cdot \frac{\sin n\left(\frac{\beta}{2} + \frac{\pi d}{\lambda}\right)}{n \sin\left(\frac{\beta}{2} + \frac{\pi d}{\lambda}\right)} \quad . \quad . \quad . \quad (61)$$

In the limiting case, as $n \to \infty$ and the spread of the antenna system remains constant, i.e. $n\left(\frac{\beta}{2} + \frac{\pi d}{\lambda}\right) = \frac{\beta_0}{2} + \frac{\pi d_0}{\lambda}$, equation (61) becomes

$$F_0 = f_0 \frac{\sin\left(\frac{\beta_0}{2} + \frac{\pi d_0}{\lambda}\right)}{\left(\frac{\beta_0}{2} + \frac{\pi d_0}{\lambda}\right)} \quad . \quad . \quad . \quad (62)$$

As before, the current is the sum of the currents in all the portions of the antenna system irrespective of phase.

In the special case, when $n = 2$, we have

$$F_0 = f_0 \cos\left(\frac{\beta}{2} + \frac{\pi d}{\lambda}\right) \quad . \quad . \quad . \quad (63)$$

(8) Antennæ Having Symmetrical Radiation About a Vertical Axis.

(*a*) *General Case.*

If the vertical axis of symmetry pass through a point O, in order to keep rigorous symmetry the antenna system must be uniformly distributed around the circumference of one or more horizontal circles having O as centre. The individual antennæ on each circle must be identical and must carry the same amplitude of current. Symmetry can still be kept if the currents in the antennæ are not in the same phase, but in this

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case the phase must vary uniformly round the circumference of each circle, and the total phase difference in a complete revolution must be a multiple of 2π, otherwise a discontinuity would occur.

The above conditions are necessary for exact symmetry. It may be possible to devise systems which do not conform with the above conditions, and yet give fairly correct symmetry, but it will probably be found that they approximate in some way to these conditions.

Let us consider the general case of n identical antennæ, uniformly distributed along the circumference of a circle of radius R (Fig. 11). Since the radiation is symmetrical about the vertical axis through the centre O, it is only necessary to consider the radiation distribution in a single vertical plane. Let it be OX. We shall take as the basis of phase the current in the antenna M, so that the phase of the current in the antenna A is $(n\phi)$ in advance of that in the antenna M. In order to preserve symmetry, n must be an integer.

The phase of the radiation in the vertical plane through OX at an angle of θ with the vertical from

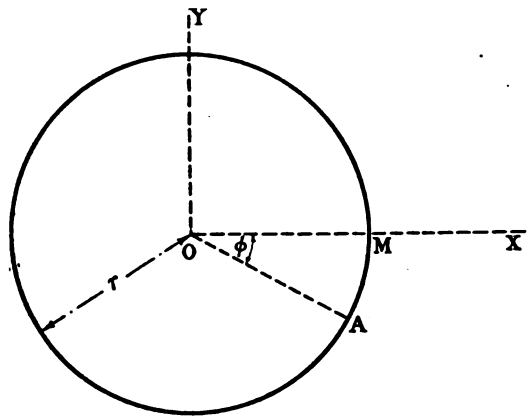

FIG. 11.

the antenna A will be in advance of that from an imaginary antenna at O, excited in the same phase as the antenna M, by an angle $(n\phi + \delta \cos \phi)$

where
$$\delta = \frac{2\pi R}{\lambda} \sin \theta \quad . \quad . \quad . \quad . \quad (64)$$

If $(f + jf')$ is the radiation distribution of each individual antenna, the radiation distribution $(F + jF')$ for the whole circle will be given by

$$F + jF' = \int_0^{2\pi} \frac{1}{2\pi}(f + jf')\{\cos(n\phi + \delta \cos \phi) + j \sin(n\phi + \delta \cos \phi)\}d\phi \quad . \quad (65)$$

As in section (7), the current is not measured at a single point but is taken to be the sum (irrespective of phase) of all the currents in the individual antennæ. The current in each of these is measured at the point which makes their radiation distribution $(f + jf')$.

The integral of equation (65) can be solved in certain special cases. In order to obtain as general a case as possible, let us assume that f and f' can each be expanded in Fourier series with the angle ϕ as variable.

$$f + jf' = \sum_m \{(A_m + jA'_m) \sin(m\phi) + (B_m + jB'_m) \cos m\phi\} \quad : \quad (66)$$

where m is an integer and A_m, A'_m, B_m, B'_m are functions of θ.

Equation (65) can be shown to reduce, when $m > n$, to

$$F + jF' = \tfrac{1}{2}\sum_m (-1)^{\frac{1}{2}(m-n+1)}\Big[(A_m + jA'_m)\{J_{m-n}(\delta) + (-1)^{n+1}J_{m+n}(\delta)\} + (B'_m - jB_m)\{J_{m-n}(\delta) - (-1)^{n+1}J_{m+n}(\delta)\}\Big] \quad . \quad . \quad . \quad (67)$$

when $(m + n)$ is odd, and to

$$F + jF' = \tfrac{1}{2}\sum_m (-1)^{\frac{1}{2}(m-n)}\Big[(B_m + jB'_m)\{J_{m-n}(\delta) + (-1)^{n}J_{m+n}(\delta)\} - (A'_m - jA_m)\{J_{m-n}(\delta) - (-1)^{n}J_{m+n}(\delta)\}\Big] \quad . \quad . \quad . \quad (68)$$

when $(m - n)$ is even, where J_r is a Bessel function of the first kind, tables of which are readily available.

Equations (67) and (68) only hold if $m > n$. When $m < n$ it is necessary to put m for n and n for m in these equations without changing any of the signs.

These general formulæ simplify considerably in a large number of special cases, some of which will now be considered.

(*b*) *Circle of Vertical Antennæ all Excited in the Same Phase.*

In this case $m = n = 0$, and from equation (7) $f'_\theta = f_\phi = f'_\phi = 0$, so that the radiation distribution of each individual antenna is a single function of the angle θ (this function is not complex). Let it be f_θ. The radiation distribution of the whole circular system will then be given by

$$\left.\begin{aligned} F_\theta + jF'_\theta &= f_\theta J_0(\delta) \\ F_\phi + jF'_\phi &= 0 \end{aligned}\right\} \quad . \quad . \quad . \quad (69)$$

When the height of the individual antenna is small $f_\theta = \delta l \sin \theta$ [see equation (3)].

It may be noted that the number of antennæ in the circle does not enter into equation (69). That is so because, as in section (7), the total current is measured and not the current in an individual antenna. This convention will be adopted in all the following cases.

(*c*) *Circle of Small Vertical Radial Coils Excited in the Same Phase.*

If the vector representing the coil at M [see section 2 (*b*)] is along the Y axis (Fig. 11), then equation (11) gives

$$f_\theta = 0, \quad f_\phi = 0, \quad f'_\theta = -(2\pi/\lambda) \cos \phi \, \delta A$$

and
$$f'_\phi = -(2\pi/\lambda) \cos \theta \sin \phi \delta A$$

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Hence, the radiation distribution of the whole circle is given by

$$\left.\begin{aligned} F_\theta + jF'_\theta &= j\frac{2\pi\delta A}{\lambda}J_1(\delta) \\ F_\phi + jF'_\phi &= 0 \end{aligned}\right\} \quad . \quad . \quad . \quad (70)$$

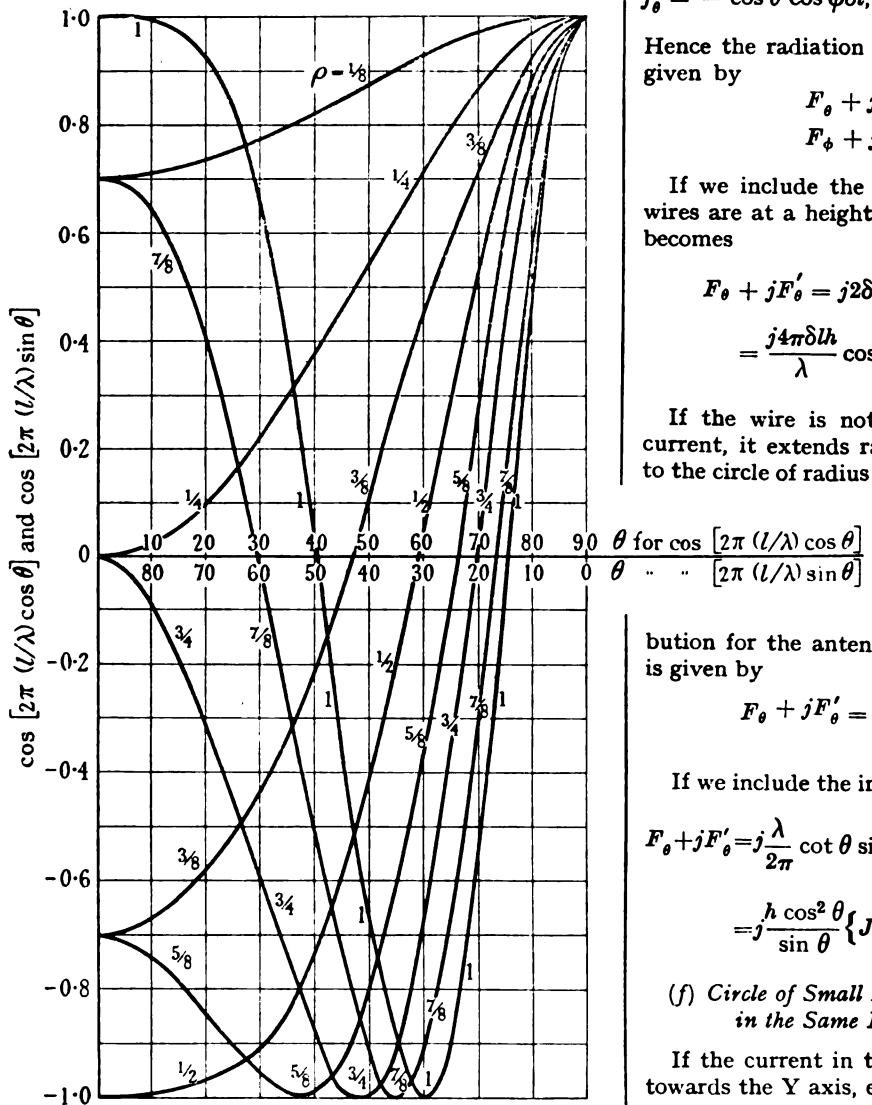

FIG. 12.—Curves of $\cos\{2\pi(l/\lambda)\sin\theta\}$ and $\cos\{2\pi(l/\lambda)\cos\theta\}$ for various values of l/λ.

(d) Circle of Small Vertical Tangential Coils Excited in the Same Phase.

If the vector representing the coil at M is radially outwards, then equation (10) gives

$$f_\theta = 0, \quad f_\phi = 0, \quad f'_\theta = (2\pi/\lambda)\sin\phi\delta A,$$
$$f'_\phi = -(2\pi/\lambda)\cos\theta\cos\phi\delta A$$

Hence the radiation distribution for the whole circle is given by

$$\left.\begin{aligned} F_\theta + jF'_\theta &= 0 \\ F_\phi + jF'_\phi &= j\frac{2\pi\cos\theta\delta A}{\lambda}J_1(\delta) \end{aligned}\right\} \quad . \quad . \quad (71)$$

(e) Circle of Small Horizontal Radial Wires Excited in the Same Phase.

If the current is flowing radially outwards, equation (8) gives

$$f_\theta = -\cos\theta\cos\phi\delta l, \quad f_\phi = -\sin\phi\delta l, \quad f'_\theta = 0, \quad f'_\phi = 0$$

Hence the radiation distribution for the whole circle is given by

$$\left.\begin{aligned} F_\theta + jF'_\theta &= \delta l\cos\theta J_1(\delta) \\ F_\phi + jF'_\phi &= 0 \end{aligned}\right\} \quad . \quad . \quad . \quad (72)$$

If we include the image in a perfect earth, and the wires are at a height h above the ground, equation (72) becomes

$$F_\theta + jF'_\theta = j2\delta l\cos\theta\sin\left(\frac{2\pi h\cos\theta}{\lambda}\right)J_1(\delta) \quad . \quad (73)$$

$$= \frac{j4\pi\delta lh}{\lambda}\cos^2\theta J_1(\delta) \text{ if } (h/\lambda) \text{ is small} \quad . \quad . \quad (74)$$

If the wire is not small, but still carrying uniform current, it extends radially from the circle of radius r_1 to the circle of radius $r_2 (r_2 > r_1)$, and the radiation distribution for the antenna system alone without its image is given by

$$F_\theta + jF'_\theta = \frac{\lambda}{2\pi}\cot\theta\{J_0(\delta_1) - J_0(\delta_2)\} \quad . \quad (75)$$

If we include the image in a perfect earth this becomes

$$F_\theta + jF'_\theta = j\frac{\lambda}{2\pi}\cot\theta\sin\left(\frac{2\pi h\cos\theta}{\lambda}\right)\{J_0(\delta_1) - J_0(\delta_2)\} \quad (76)$$

$$= j\frac{h\cos^2\theta}{\sin\theta}\{J_0(\delta_1) - J_0(\delta_2)\} \text{ if } (h/\lambda) \text{ is small} \quad (77)$$

(f) Circle of Small Horizontal Tangential Wires Excited in the Same Phase.

If the current in the vector at M (Fig. 11) is flowing towards the Y axis, equation (9) gives

$$f_\theta = -\cos\theta\sin\phi\delta l, \quad f_\phi = \cos\phi\delta l, \quad f'_\theta = 0, \quad f'_\phi = 0$$

Hence the radiation distribution due to the whole circle without its image is given by

$$\left.\begin{aligned} F_\theta + jF'_\theta &= 0 \\ F_\phi + jF'_\phi &= -\delta lJ_1(\delta) \end{aligned}\right\} \quad . \quad . \quad . \quad (78)$$

If we include the image in a perfect earth, this becomes

$$F_\phi + jF'_\phi = -j2\delta l\sin\left(\frac{2\pi h\cos\theta}{\lambda}\right)J_1(\delta) \quad . \quad . \quad . \quad (79)$$

$$= -j\frac{4\pi h\delta l}{\lambda}\cos\theta J_1(\delta)$$

when (h/λ) is small . (80)

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(8) EXAMPLE.

A case which occurs in engineering practice is that of the beam antenna. As an example, suppose a beam system to consist of n equal vertical antennæ carrying equal currents and spaced a distance d apart. Let us further suppose that each antenna consists of m positive

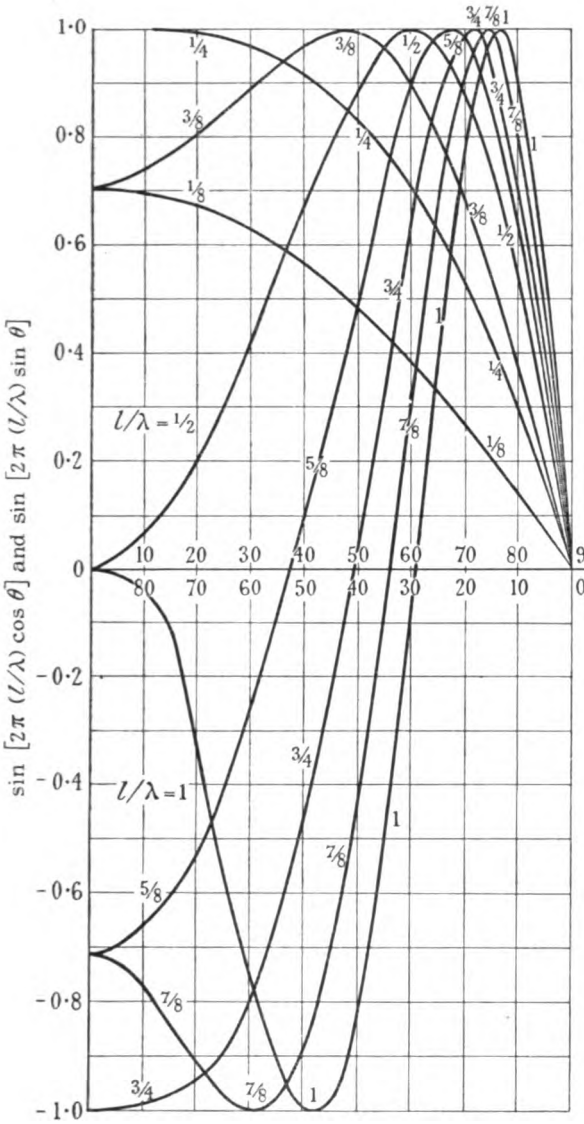

FIG. 13.—Curves of $\sin\{2\pi(l/\lambda)\sin\theta\}$ and $\sin\{2\pi(l/\lambda)\cos\theta\}$ for various value of l/λ.

half-waves of current superposed one above the other, the negative half-waves being suppressed by the Franklin or other system.

From equation (46) a half-wave vertical antenna has a radiation distribution given by

$$f_0 + jf_0' = \frac{\lambda}{2\pi \sin\theta}\left\{\cos(\pi\cos\theta) + 1 + j\sin(\pi\cos\theta)\right\} \quad . \quad (81)$$

The resultant of this is given by

$$f_0 = \frac{\lambda}{\pi\sin\theta}\sqrt{\left[\tfrac{1}{2}\left\{1+\cos(\pi\cos\theta)\right\}\right]} \quad . \quad (82)$$

By means of equations (58) and (61) we can superpose m of these half-wave elements on the top of one another by putting $x_r = y_r = 0$ and $z_r = \frac{1}{2}\lambda$ in equation (82), and substitute the value of d obtained from equation (61), so that the radiation distribution of each antenna without the image in the ground is given by

$$F_0 = \frac{f_0}{m}\cdot\frac{\sin(m\pi\cos\frac{1}{2}\theta)}{\sin(\pi\cos\frac{1}{2}\theta)} \quad . \quad . \quad . \quad (83)$$

The radiation distribution for the whole antenna array assumed spread along the X axis without the image in the ground is, from equations (58 and (61),

$$\frac{F_0}{n}\cdot\frac{\sin(n\pi d\cos\phi\sin\theta/\lambda)}{\sin(\pi d\cos\phi\sin\theta/\lambda)} \quad . \quad . \quad . \quad (84)$$

If the ground is of perfect conductivity, since the image in this case is identical with the antenna system itself, the current flowing in the same direction in both, it is only necessary to replace m by $2m$ in order to obtain the radiation distribution for the combination of the antenna and its image.

If the ground is not perfect, the total result can be obtained from equation (34), making use of the curves given in Figs. 4 to 9.

In practice the individual antennæ forming an array are sometimes so complicated that it is not practical to reduce the radiation distribution to the form of an equation. It is best in this case to obtain the radiation distribution of the individual antennæ graphically, and apply equations (58) and (61) to the result in order to find the radiation distribution for the whole array.

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THE RADIATION DISTRIBUTION OF ANTENNÆ IN VERTICAL PLANES.*

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(*Paper first received 16th November, 1929, and in final form 22nd February, 1930.*)

SUMMARY.

The radiation distribution of an antenna in a vertical plane was obtained by measuring in an aeroplane the strength of a received signal from an excited antenna on the ground. The position of the aeroplane relative to the antenna was obtained by means of a theodolite on the ground, the signal strength being recorded on a cinematograph film.

Although the ground site appeared very good, it was found that spurious radiations were obtained and these were probably the cause of the discrepancies found between the theoretical and the experimental results. The measurements were made on a horizontal antenna at several frequencies. The reason for this particular arrangement is that it should be possible to separate the effect of the reflection of the antenna in the earth from that of the antenna itself, and thereby obtain a double result from a single measurement.

The results showed very definite maxima and minima and the position of these could be ascertained accurately within a few degrees, but the difficulties of the experiment prevented an accurate value of the field strength being obtained. Substantial agreement was obtained between theory and experiment, the discrepancies increasing with the frequency.

The radiation distribution was also obtained for one of the beam stations erected by the Marconi Co. Owing to the sharpness of the beam, considerable experimental error was liable to occur. Nevertheless, substantial agreement between theory and experiment was also obtained in this case.

(1) INTRODUCTION.

A considerable amount of experimental data has been obtained on the radiation distribution from antennæ in a horizontal plane, and all the evidence strongly shows that the theoretical and the experimental results are in very close agreement, particularly for short-wave stations. On longer waves the mast system is usually a much more important part of the whole erection and it is probable that the discrepancies, which are often found where steel masts are used, are due to induced currents in the masts which are not allowed for in the calculations. There are very few cases, with the conspicuous exception of the Beverage antenna, in which the radiation distribution in a horizontal plane depends on the distribution of current along a horizontal wire, and it was with the double purpose of investigating whether such a radiation distribution could be calculated and of examining the effect of the reflection in the earth that the experiments described in this paper were undertaken.

The distribution of current along a transmitting antenna has been discussed theoretically by Macdonald † and Abraham ‡ with substantial disagreement as to the size of the first loop near the end of the wire. Macdonald makes the distance to the first current node as low as $\lambda/5$ in some cases, while Abraham obtains $\lambda/4{\cdot}2$ for this distance. Abraham considered the case of a thin ellipsoid, and Lord Rayleigh § has shown that, in spite of the tapering of the wire which this assumption involves, the result is applicable to a thin circular wire. The experimental evidence is to some extent conflicting, but most of the results date from the time when spark sets were the only available source of oscillations. Few experiments have been carried out on current distribution since the advent of the wireless valve. The author ‖ recently published some results on an 11-metre vertical antenna over a range of wavelengths from 15 m to 800 m and found agreement with Abraham's theory to an order of accuracy of some 5 per cent. This means that the accuracy was not sufficient to discriminate between Abraham's theory ($\lambda/4{\cdot}2$) and the simple sinusoidal theory ($\lambda/4$). The current was measured by ammeters, the capacity and self-inductance effects of which were accurately considered. C. R. Englund ¶ also obtained agreement with the simple theory, by finding the natural frequency of straight rods. It would therefore appear that, to a sufficient degree of accuracy, one can assume that the current distribution follows a simple sinusoidal law with small corrections for the end effects (see Appendix 1), the distance between nodes and antinodes being $\lambda/4$.

There is some further evidence from certain field-strength measurements that the simple sinusoidal law of current distribution holds and that, moreover, there is no appreciable change in phase along the antenna. One of these confirmatory results was recently given by P. P. Eckersley, T. L. Eckersley and H. L. Kirke,** who showed that the field strength on the ground from the half-wave vertical antenna was double that of a quarter-wave antenna to an accuracy of 5 per cent (well within the limits of experimental error), showing that both quarter wavelengths of the antenna were radiating equally and were in the same phase. Further evidence was obtained from investigations carried out by G. H. Munro in connection with attenuation experiments undertaken by R. H. Barfield.†† In order to obtain the effective height of an antenna, Munro suggested using a "T" antenna and increasing the length of the horizontal portions until the signal received from a constant source

* Reprinted from *Journal I.E.E.*, 1930, vol. 68, p. 1191.

† See Reference (1). ‡ *Ibid.*, (2). § *Ibid.*, (3).
‖ *Ibid.*, (4). ¶ *Ibid.*, (5). ** *Ibid.*, (6).
†† See *Journal I.E.E.*, 1928, vol. 66, p. 215; also *Proceedings of the Wireless Section*, 1928, vol. 3, p. 14.

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remained substantially constant, when the effective height could be assumed to be equal to the actual height. The variation of effective height, calculated by the author from the simple sinusoidal current distribution theory,* is shown in Fig. 1. The crosses represent the experimental values obtained by Munro. It is interesting to note how closely the experimental and theoretical results agree.

There is considerable evidence, therefore, in favour of the simple sinusoidal theory of current distribution, and this is the theory used in this paper in comparing the theoretical and experimental radiation distribution diagrams. A number of general formulæ for this have been obtained by the author.†

The other factor that enters into the theoretical consideration of this problem is the effect of the earth and its imperfect conductivity. A theory has been put forward by T. L. Eckersley‡ and amplified by the author § (Appendix 2), but there has not yet been any satisfactory experimental confirmation of this theory.

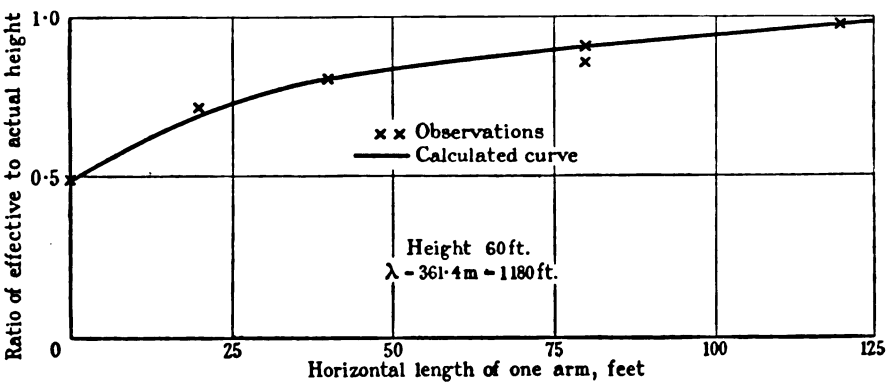

FIG. 1.—Effective height of " T " antenna.

(2) APPARATUS AND PROCEDURE.

For the purpose of these experiments the Air Ministry kindly placed an aeroplane at the disposal of the Radio Research Board. It was fitted with a " T " antenna having a horizontal portion 27 ft. long spread between two uprights 4 ft. high at each wing tip and supported at the centre. A supersonic heterodyne receiver was used. The intermediate- and low-frequency amplifiers were standard R.A.F. equipment. The low-frequency output was measured on a valve voltmeter using anode rectification. The detecting instrument consisted of a double-pivot microammeter from which the pointer had been removed, a small mirror being substituted. A small 2-volt lamp was focused on the mirror and reflected on to a cinematograph film kept moving by means of a gramophone motor at the rate of about 30 ft. per hour. The sensitivity for full-scale deflection was 50 μA. The whole was, of course, completely screened from external light. In order that the observer in the aeroplane might ensure that the spot of light remained on the film, the beam was intercepted by a semi-silvered mirror which reflected some of the light on to a red ground-glass screen on which the observer could see the deflection of the instrument. The galvanometer was fitted with shunts of several values so that, by means of a switch, it could be made more or less sensitive, and it was the duty of the observer in the aeroplane to operate the switch in order to keep the galvanometer at some suitable deflection.

A better, method but one which would be slightly more cumbersome to apply satisfactorily, would have been to insert a potentiometer or attenuator device in the intermediate-frequency amplifier (as in the field-strength measuring set due to Friis and Bruce *) and thus control the galvanometer deflection, but it was not considered at first that the experimental accuracy, especially in locating the aeroplane over the whole flight, would warrant this extra encumbrance. The experience gained, however, showed that this was an error of judgment and that more reliable field-strength measurements would have been useful.

The receiving set was calibrated on the ground by means of a small local oscillator in which the antenna current could be varied and measured over a range of 100 to 1.

The transmitting antenna was spread between two wooden masts. The ground was flat and free from obstructions, and the site appeared to be very suitable and likely to approach fairly closely to the ideal case. There were a number of buried metal pipes and electric cables in the neighbourhood, as well as an earth which consisted of long aluminium strips embedded in the ground and spreading out radially from the centre of the antenna.

The aeroplane was piloted to glide from a height of 5 000 ft. down to 500 ft. from one side of the antenna to the other in such a direction that the two masts and the aeroplane remained in the same vertical plane. In this way it was possible to ascertain the position of the aeroplane from the horizontal and vertical angles obtained with a single theodolite. The theodolite was situated at a distance of some 3 miles from the station in a direction perpendicular to the direction of flight. At this distance three men were able after some practice to follow accurately the course of the flight and take readings on the theodolite every 10 seconds.

It is no easy matter for a pilot to keep his aeroplane

* See Appendix 1 of Reference (4).
† See Reference (7
‡ *Ibid.*, (8).
§ *Ibid.*, (9).
* See Reference (10).

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flying accurately in one plane for, when a cross wind is blowing, as was almost invariably the case, the nose of the aeroplane may point 30° or more away from the true direction of flight. In order to give the pilot an indication by which to correct his direction of flight his telephones were connected in series with those of the observer in the aeroplane so that he could hear the modulated signal transmitted from the ground station. This signal was stopped for the space of a dot or a dash. A dot meant that he should turn one way and a dash the other way: if he were much out of his course a double signal would be sent. In this way the pilot could adjust his course correctly within a few degrees.

The directional signals were sent every 10 seconds in synchronism with the readings taken on the theodolite, so that the cinematograph film on which the signal

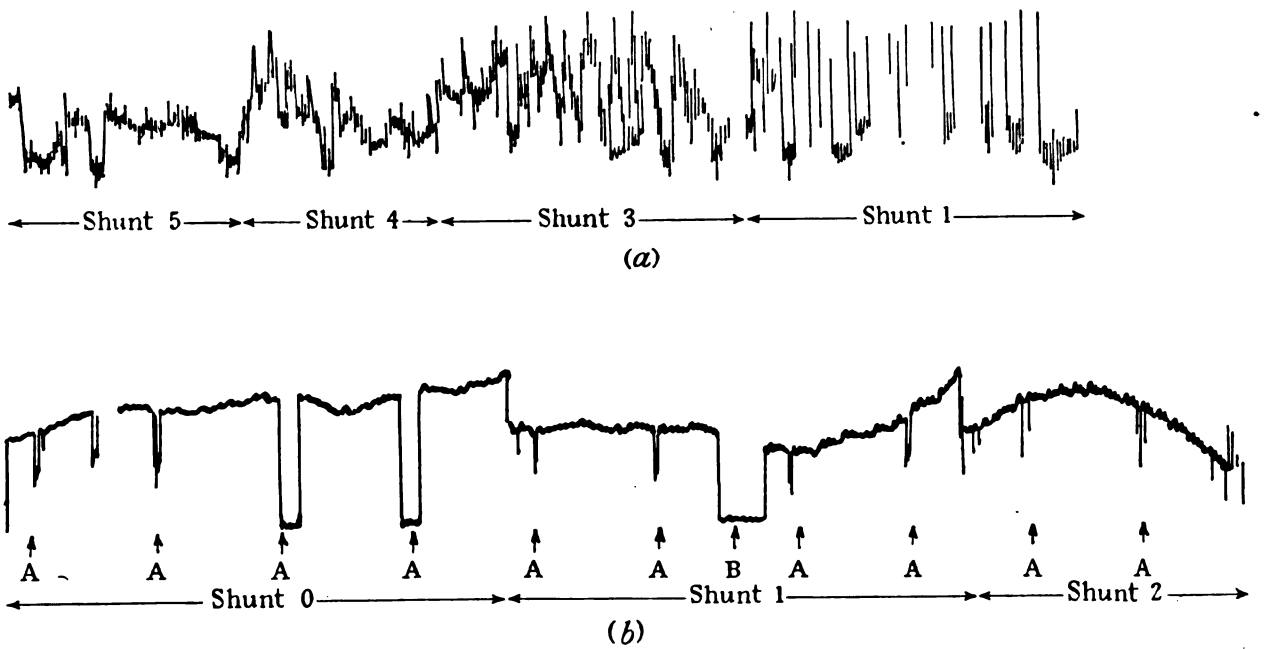

FIG. 2.

strength was recorded had a time scale automatically recorded on it [see Fig. 2 (*b*)]. Every minute a special signal was sent.

It may be realized that the experiments might have been inversed, i.e. a transmitter might have been installed in the aeroplane and the signal received on the ground. Such an arrangement would have permitted the use of a more elaborate signal-measuring set, but the directional indications to the pilot by means of the scheme described would not have been possible. Such indications would have necessitated communication on another frequency and probably the erection of a second antenna on the aeroplane. This was not considered advisable, for the radiation distribution of the aeroplane antenna is an unknown quantity and it seemed dangerous to distort it unnecessarily.

Baldus and Hase,* while investigating the radiation distribution of a trailing antenna in an aeroplane, employed transmissions from the aeroplane. They did not use any signal to direct the pilot, who found it almost impossible to fly correctly. The method employed in this case to locate the aeroplane was to take photographs of the ground and read the height. This method is liable to considerable error, first because the aeroplane may not be flying horizontally and a correction has to be applied from the readings of the levels, and secondly because the height is not easily measured. An accurate pressure altimeter was installed in the aeroplane used by the author. On one occasion when it was checked against the theodolite reading it showed very good agreement, but when the temperature correction was applied considerable divergence was found.

The preliminary experiments consisted mainly in adjusting the sensitivity of the receiver. The engine was found to be insufficiently screened. The comparison of the film readings shown in Fig. 2 are interesting. Fig. 2 (*a*) shows the type of field strength received when the aeroplane was first tried. After investigating the sources of radiation on the engine and screening a 2-in. length of exposed lead, the signal received was steadied to that shown in Fig. 2 (*b*). To compare Figs. 2 (*a*) and 2 (*b*), it is necessary to compare those parts for which the receiving set was of the same sensitivity. As explained above, variation of sensitivity was obtained by means of a series of shunts numbered from 0 to 6 (shunt 0 being the condition for which the set was most sensitive). Changes in shunt can be seen by the sudden changes in the signal strength, which are readily noticed in Fig. 2 (*b*). The sudden dips marked A in Fig. 2 (*b*) are the directional signals explained above, and the long dash seen at B is the distinguishing signal sent every

* See Reference (11).

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minute. A more quantitative investigation of shielding in aeroplanes has been carried out recently by Drake.*

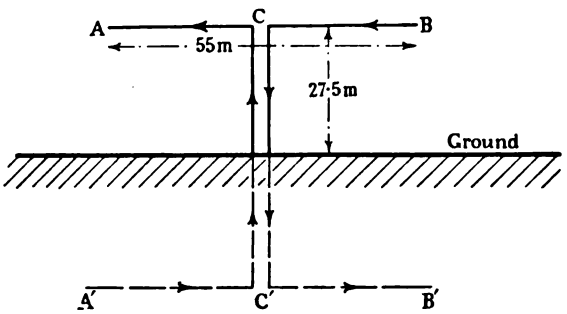

FIG. 3.

The antenna investigated was a horizontal one (Fig. 3) supported by wooden masts 150 m apart. It was excited by means of a push-pull oscillator, and the currents in the two leads were adjusted to equality at the foot of the antenna. The vertical leads were spaced 6 in. apart and the radiation from them was calculated to be quite negligible. This type of antenna is particularly interesting in this experiment because the whole of its radiating portion is at a constant height above the ground. Suppose an element P of the antenna together with its image has a radiation distribution of $I_p f_r$, where I_p is the current, then the function f_r will be the same for all the other elements of the antenna. If the whole antenna AB, considered in free space without its image, has a radiation distribution of f_a, then the total radiation distribution of the complete antenna with its image will be $f_a \times f_r$. By investigating such an antenna it should be possible to gain information regarding both f_a and f_r separately.

From the theory for the effect of the earth, mentioned above, it can be shown that, if $f_r = 0$ for a certain angle to the vertical in the case of a perfect earth, f_r will reach a minimum at a larger angle to the vertical when the earth is imperfect, and, moreover, f_r will never become zero in this case.

The variation of field strength recorded on the cinematograph film can only be construed into radiation distribution on two assumptions, viz.

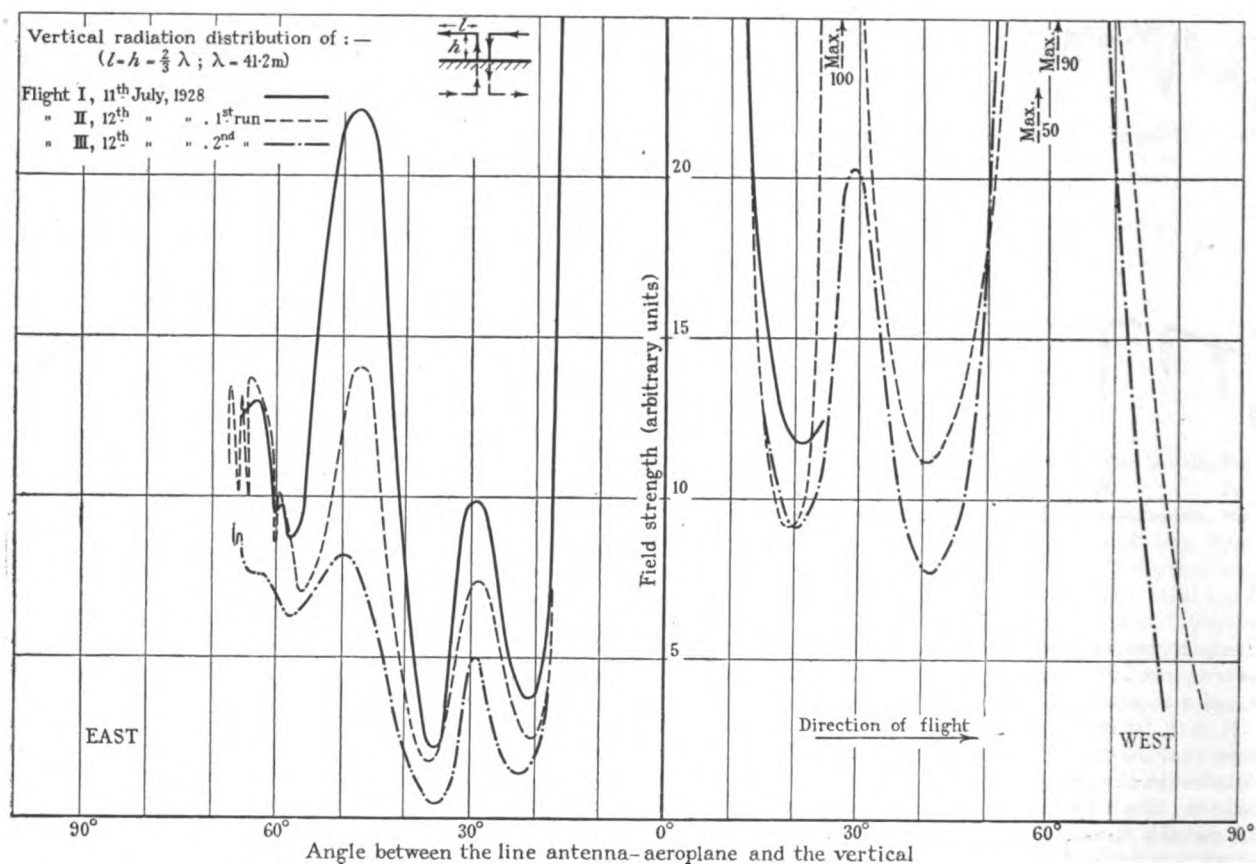

FIG. 4.

(a) That the field strength varies inversely as the distance, and

(b) That the radiation distribution of the aeroplane antenna is sinusoidal.

The first assumption would appear to be incorrect for

* See Reference (12).

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the ground ray, on account of attenuation, but along any other line there is no reason to suppose that this would be so. In fact, experimental evidence in support of the assumption is given below.

Some experiments described below were carried out in an endeavour to verify or alter the second assumption, but the results did not justify the use of any more definite law than the simple sinusoidal distribution, which is based on the fact that the aeroplane antenna was invariably small compared with the wavelength. It will be seen that this gives rise to considerable difficulty when the aeroplane is vertically above the transmitting antenna, for, in that case, if the signal strength in the aeroplane is not zero, the calculated strength of the radiation from the transmitter will be infinite. In fact, for all angles within some 7° of the vertical the radiation distribution will be very inaccurate indeed and will consequently be omitted in the radiation distribution diagrams found experimentally. These diagrams are not plotted in the usual polar co-ordinates but in cartesian co-ordinates, so as to make the discrepancies more clearly visible. The angles in degrees measured from the vertical are plotted along the abscissæ, while the field strengths are plotted along the ordinates. Since the absolute values of field strengths were not determined the units used are absolutely arbitrary.

(3) Experimental Results on Simple Antennæ.

The first tests were carried out at a wavelength of 41·1 m ($h = \frac{2}{3}\lambda$) on a horizontal antenna as shown in Fig. 3, and the results of three flights all in the same direction are shown in Fig. 4.

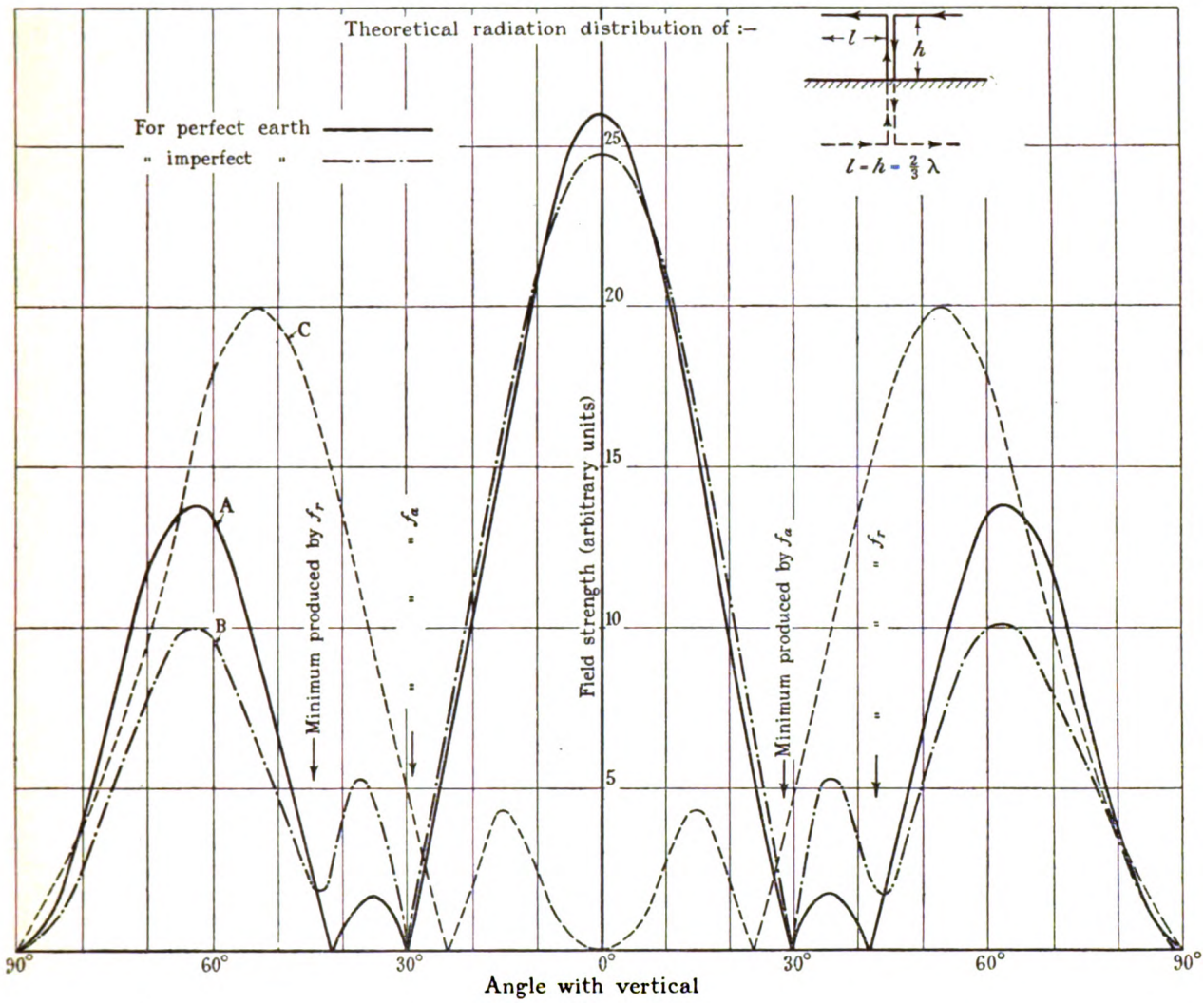

Fig. 5.

In analysing the experimental results it is essential to endeavour to discriminate between the radiation distribution due to the antenna alone (f_a) and that due to the effect of reflection in the earth (f_r). The radiation distribution in the case of a perfectly conducting earth is shown by the full line in Fig. 5, curve A. The minimum at 30° is due to f_a, while that at 41·5° is due to f_r. Mr. Barfield kindly measured the conductivity of the ground

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on the site of the transmitting antenna, by observing the tilt of the ground wave. He obtained values which varied between $0{\cdot}8 \times 10^8$ and 2×10^8 electrostatic units at various frequencies, the variation being caused, he thought, by buried conductors. Assuming that the conductivity is $1{\cdot}5 \times 10^8$ electrostatic units, the radiation can be calculated as explained in Appendix 2, and gives a distribution shown by the chain curve in Fig. 5 (curve B). It will be seen that the minimum due to f_a remains unaltered, while that due to f_r does not come to zero and is shifted to a slightly larger angle with the vertical. If the end-effect correction explained in Appendix 1 be applied, the curve will not be affected

minima on either side of the antenna might be caused by some error due to synchronizing the watches. An error of 1 second might account for this, and, although the watches were thoroughly tested and were checked before and after the flights, such an error might be conceivable owing to changes of temperature in the course of the flights, which were some 2 hours in duration. It will be noted, however, that the error repeats accurately on two different days and it is difficult to imagine how such a consistent error could occur.

It was thought that some of the discrepancies between the theoretical and the experimental curves might be caused by slightly unequal excitation of the two halves

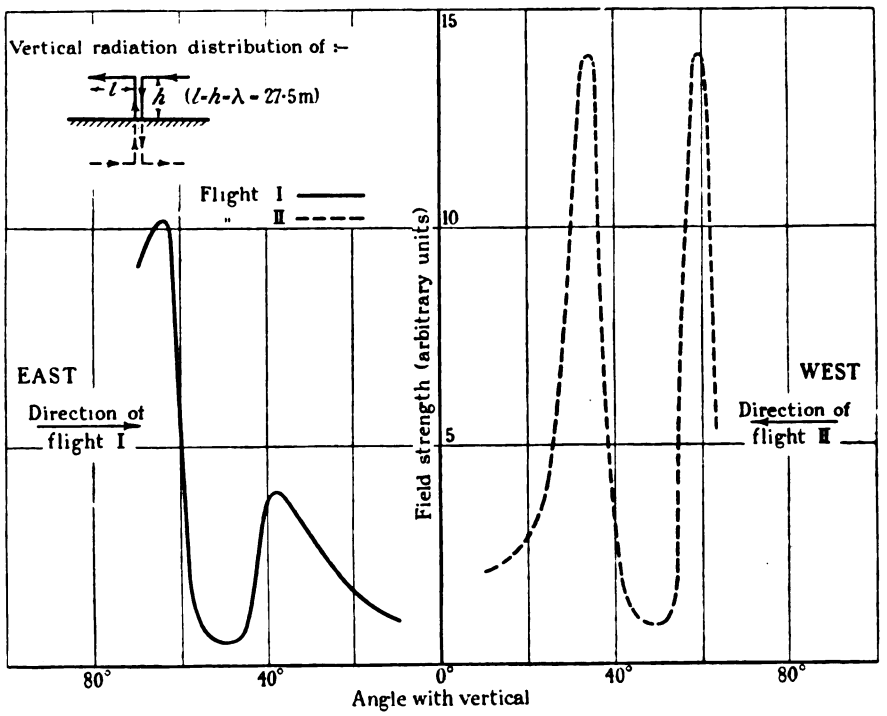

FIG. 6.

to a very great extent, but the minimum due to f_r will be shifted some 3° towards the vertical.

On examining Fig. 4 a number of points may be noted:

(*a*) The positions of the main maxima and minima repeat to an accuracy of less than 2°.

(*b*) The positions of the main maxima and minima are slightly different on the two sides of the transmitting antenna and occur at a greater angle to the vertical than the theory would lead one to expect.

(*c*) The signal strength on the east side is much weaker than that on the west.

The accuracy with which the results repeat leaves very little room to doubt that all the effects noted truly exist and are not merely due to experimental error. The asymmetry of the positions of the maxima and

of the antenna. Calculation showed, however, that this could not cause asymmetry and that its effect would only be of the second order, as it would produce a component of the radiation distribution in time quadrature with the main component. This component is shown by the broken line C in Fig. 5 and it does not appear that any satisfactory deduction can be made from it.

The minimum that should theoretically occur at 43° is the minimum produced by the interference of the antenna and its image, Experimentally it is found to occur at 36° on one side and 41° on the other, giving a mean of 38½°. If the earth were a perfect conductor this minimum would occur at 41°. It would appear, therefore, that the minimum is shifted by the imperfection of the earth by the correct amount but in the wrong direction.

The position of the other minimum, which is produced by the antenna alone, shows a discrepancy of as much as 9° between the experimental and theoretical values.

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If allowance is made for end effects, as explained in Appendix 1, this discrepancy can be reduced to 5°.

ductors, etc.—and evidence for this will be given below.

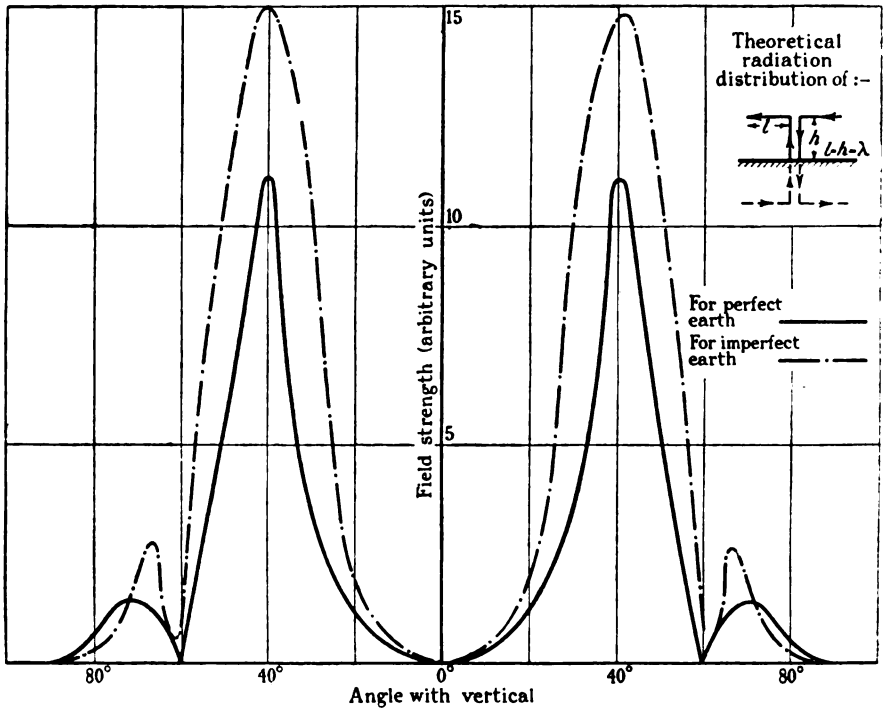

FIG. 7.

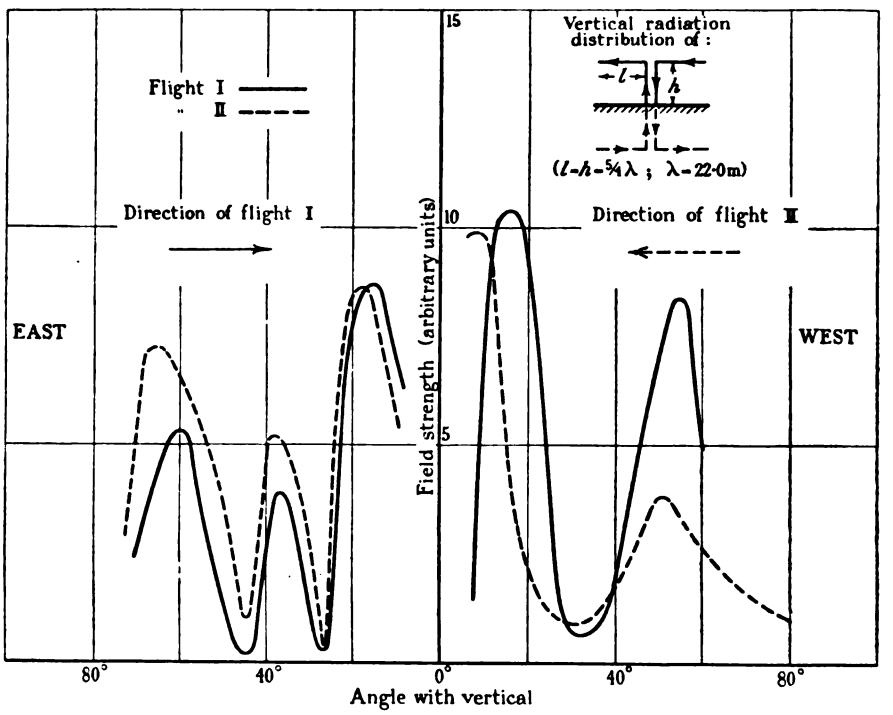

FIG. 8.

The author believes that the reason for these discrepancies may well be found in the site—buried con-

The next experiments were carried out on the same antenna with a wavelength of 27·5 m ($h = \lambda$). The

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experimental results are shown in Fig. 6. In this case the flights were carried out over the antenna from both directions in the hope of eliminating the effect of the directive properties of the aeroplane itself. It will be seen later that this could not give the expected results. The curves of Fig. 6 are to be compared with the theoretical curves of Fig. 7.

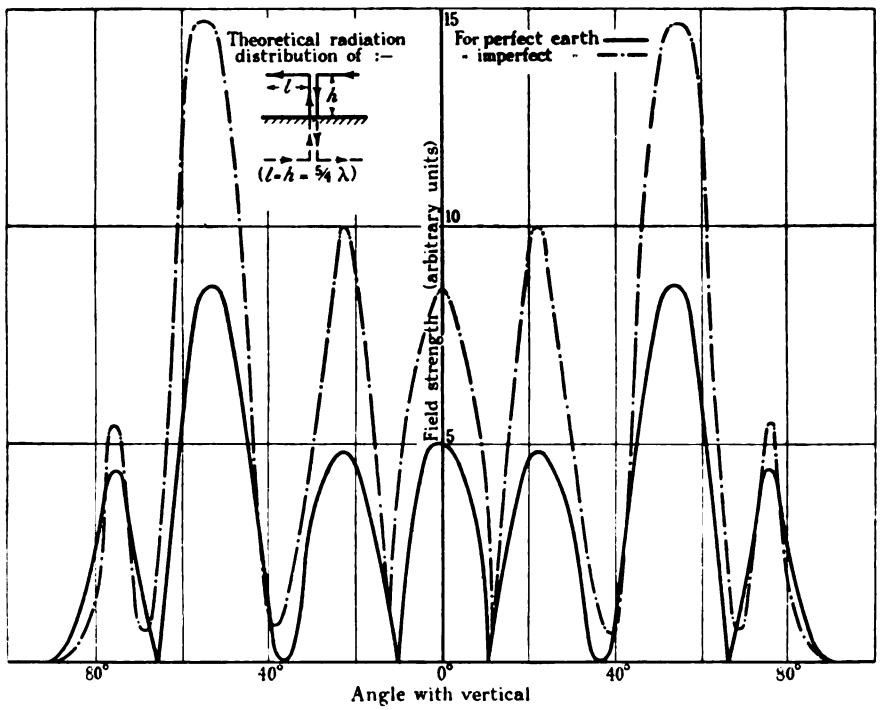

FIG. 9.

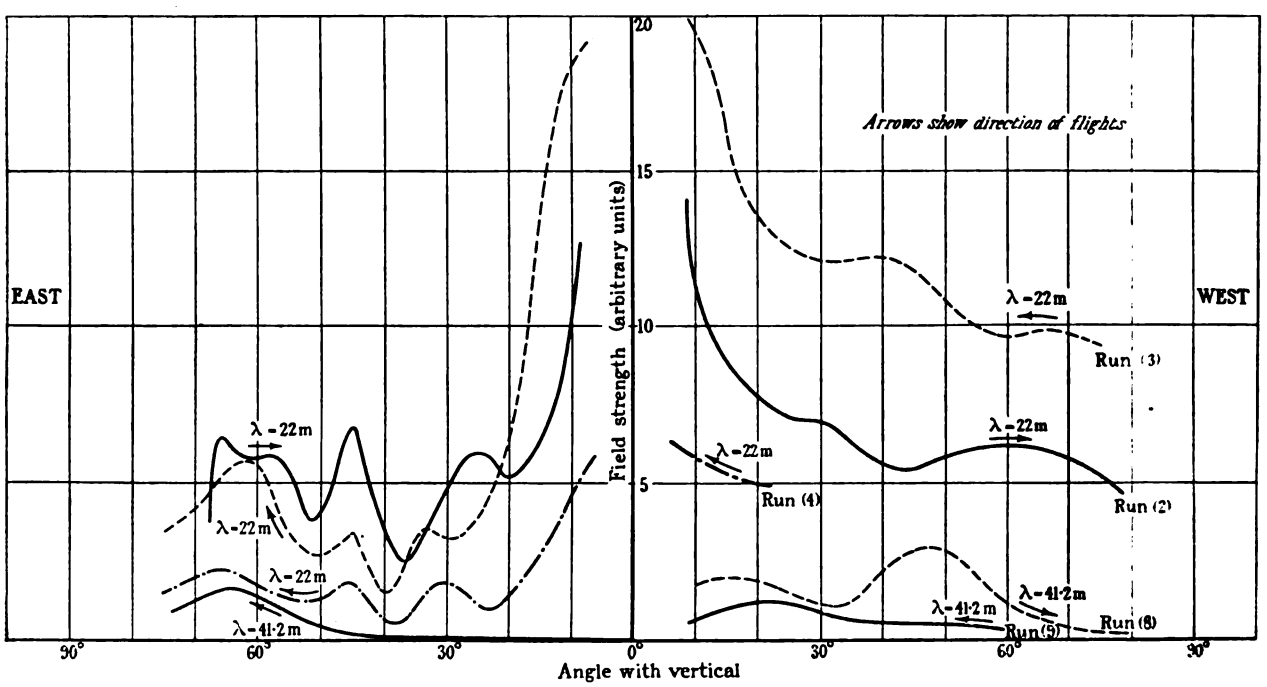

FIG. 10.—Radiation distribution with short vertical antenna.

Arrows show direction of flights

Deductions similar to those in the previous case can be made, namely, asymmetry on the two sides of the antenna and a shifting of the minima and maxima

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towards the vertical as compared with the theoretical curves.

Similar tests were made with the antenna excited at a wavelength of 22 m ($h = \frac{1}{4}\lambda$). The experimental results are shown in Fig. 8, which should be compared with the theoretical curves of Fig. 9. Here, again, deductions similar to those in the two previous cases can be made.

In order to allow for the radiation distribution of the antenna on the aeroplane, the field strength received from a small vertical antenna was analysed with the aeroplane in the same way as in the case of the horizontal antenna. By a small antenna is meant an antenna not greater than $\frac{1}{8}$th wavelength in height. In order to increase the effective height of this antenna, a horizontal portion was added in the plane perpendicular to the plane of flight of the aeroplane, so that radiation from this horizontal portion might affect the result to a negligible extent. In these conditions it was thought reasonable to assume that the radiation distribution of the transmitter would be sinusoidal. That is, if the signal strength (corrected for distance) were divided by sin θ (where θ is the angle with the vertical) the result should give the radiation distribution of the aeroplane antenna. In this, however, the experiments led to disappointment. The results obtained are shown in Fig. 10.

Two wavelengths were studied (41·2 m and 22 m). The aeroplane was flown both from east to west and from west to east, so that in some cases the engine would lie between the receiving and the radiating antenna and in the other cases it would lie on the other side of the receiving antenna, and the screening effect might be expected to be different in the two cases. On examining Fig. 10, however, it will be found that the radiation distribution diagram is substantially independent of the direction of flight of the aeroplane. Yet the curves obtained are far from uniform and the signal strength on the east side of the transmitting antenna is always much stronger than that on the west. Looking back at Figs. 4, 6 and 8 we see that this result was obtained in every case. We are finally led to the

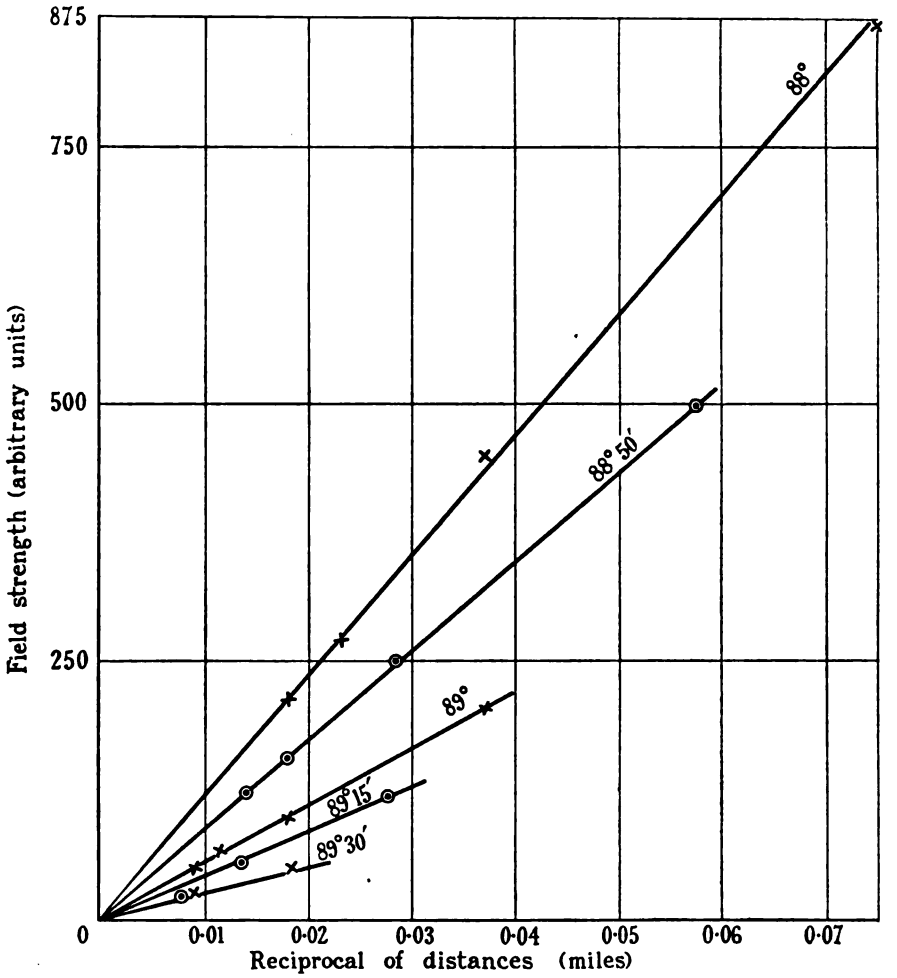

FIG. 11.—Signal strength plotted against reciprocal of distance for various inclinations to vertical. $\lambda = 185$ m.

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conclusion that the radiation distribution of the aeroplane antenna may be sinusoidal and that the irregularities noted are due to the nature of the neighbouring objects around the transmitter. This result was unexpected as the site was considered to be clear of interfering obstructions. The building in which the oscillator was housed at the foot of the antenna was made of concrete without any metal, but the back of it contained electrical power generators some 60 ft. from the foot of the antenna. There were also some conductors buried in the ground. The masts were made of wood and held up by steel-rope stays, but the masts were 500 ft. apart and the antenna, held in the centre, was only 170 ft. long. There were, moreover, no trees for some considerable distance from the site. However unexpected the result may be, there does not appear to be any way of evading the conclusion

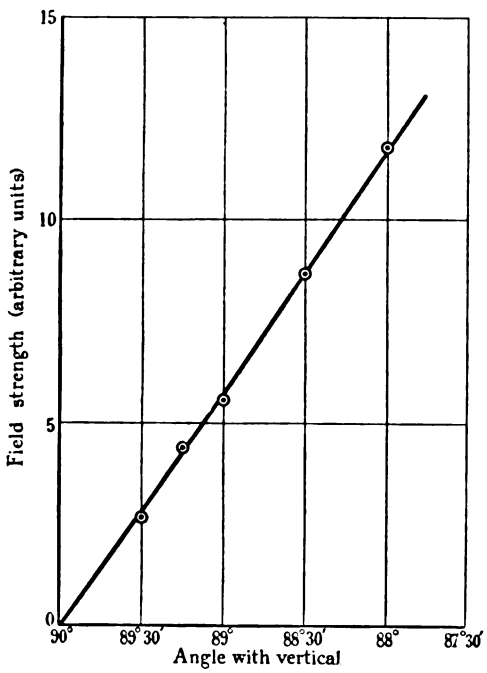

FIG. 12.—Variation of signal strength with angle to vertical. $\lambda = 185$ m.

that the variations shown in Fig. 10 were due to the ground or objects in the neighbourhood of the transmitting antenna. A further confirmation of this was obtained by finding the radiation distribution of the aeroplane antenna on the ground by rotating the aeroplane about a vertical axis. The signal strength during any one rotation was found not to vary by more than 5 per cent even at a wavelength as low as 15 m.

In the calculation of the foregoing experimental curves it has been tacitly assumed that the signal strength varies inversely with the distance. While this is known to be true for comparatively short distances on long wavelengths, it is known to be false for short waves on the ground, the law in this case being very nearly inversely as the square of the distance. The author believed that this was due to the attenuation produced by the ground, but that the inverse-distance law would hold along any radius even at very small angles of elevation from the ground. Some rough experiments carried out by the author confirmed this, but far more elaborate experiments have been carried out by D. K. Martin * of the Bell Telephone Laboratories. In these experiments the signal strength was measured at various heights above the ground, travelling horizontally for a distance as great as 120 miles from the transmitter. From these results Fig. 11 has been calculated. This shows the

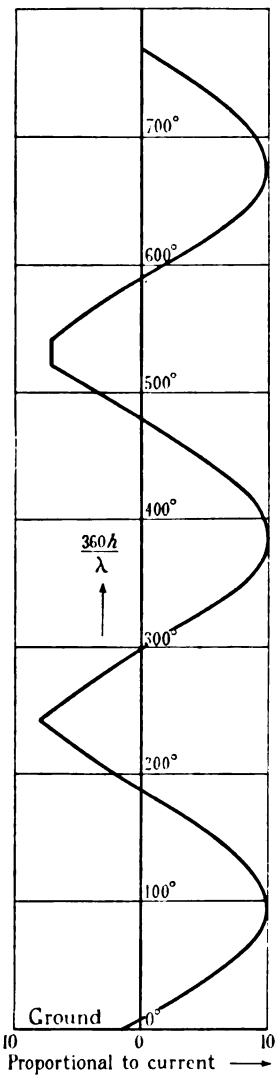

FIG. 13.—Current distribution in Australian beam antenna at Grimsby. $\lambda = 25{\cdot}9$ m.

variation of signal strength with the reciprocal of the distance along radii making various angles with the vertical. It will be noted that all the points lie on a series of straight lines to a very high degree of accuracy, thus effectively proving the inverse-distance law up to angles of $89\frac{1}{2}°$ with the vertical.

This series of curves is very interesting in another

* See Reference (13).

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respect. If we write $(\frac{1}{2}\pi - \delta)$ for θ in equation (3) of Appendix 2, and make δ very small (δ being the angle of elevation with the ground) we have

$$K_i - jK_i' = -\left\{1 - 2\delta\sqrt{(\sigma/f)} + 2j\sqrt{(\sigma/f)}\delta\right\}$$

The signal strength for small values of δ is, therefore, proportional to

$$2\delta\sqrt{(2\sigma/f)}$$

The gradient of the straight lines in Fig. 11 will be a measure of the field strength for various values of δ and, according to the above, this should be a linear relation. That this is so is well shown in Fig. 12. The gradient of this line depends on the conductivity of the ground and on the effective height and the current in the transmitting antenna. Knowing the latter, the conductivity of the ground could be deduced.

The linear relation of Fig. 12 is a very striking, though far from complete, confirmation of the equations (8) and (9) explained in Appendix 2. It is very interesting, however, in tending to show that only the constants of the ground in the immediate neighbourhood of the ground station affect the signal strength of the direct ray in aeroplane-to-ground communication.

(4) Experimental Results on a Beam Antenna.

With the co-operation of the Post Office, a number of flights were made over the beam antenna erected near Grimsby by the Marconi Co. for transmissions to Australia. The beam array consists of a large number of vertical wires spread over a span of 1 300 ft. Each wire is built on the Franklin system, with portions of current loops eliminated. The current distribution, on the simple sinusoidal theory, is shown in Fig. 13. The ordinates are measured in degrees instead of metres (a distance h corresponds to 360 h/λ degrees), to simplify the calculations. In order to obtain a theoretical radiation distribution curve the current areas are divided up into a large number of small areas having a height of not more than 45°, and the distances of their centres of gravity from the ground are read off in degrees. If A is an area and h_1 the height of its centre of gravity from the ground, the field strength in a direction making an angle θ with the

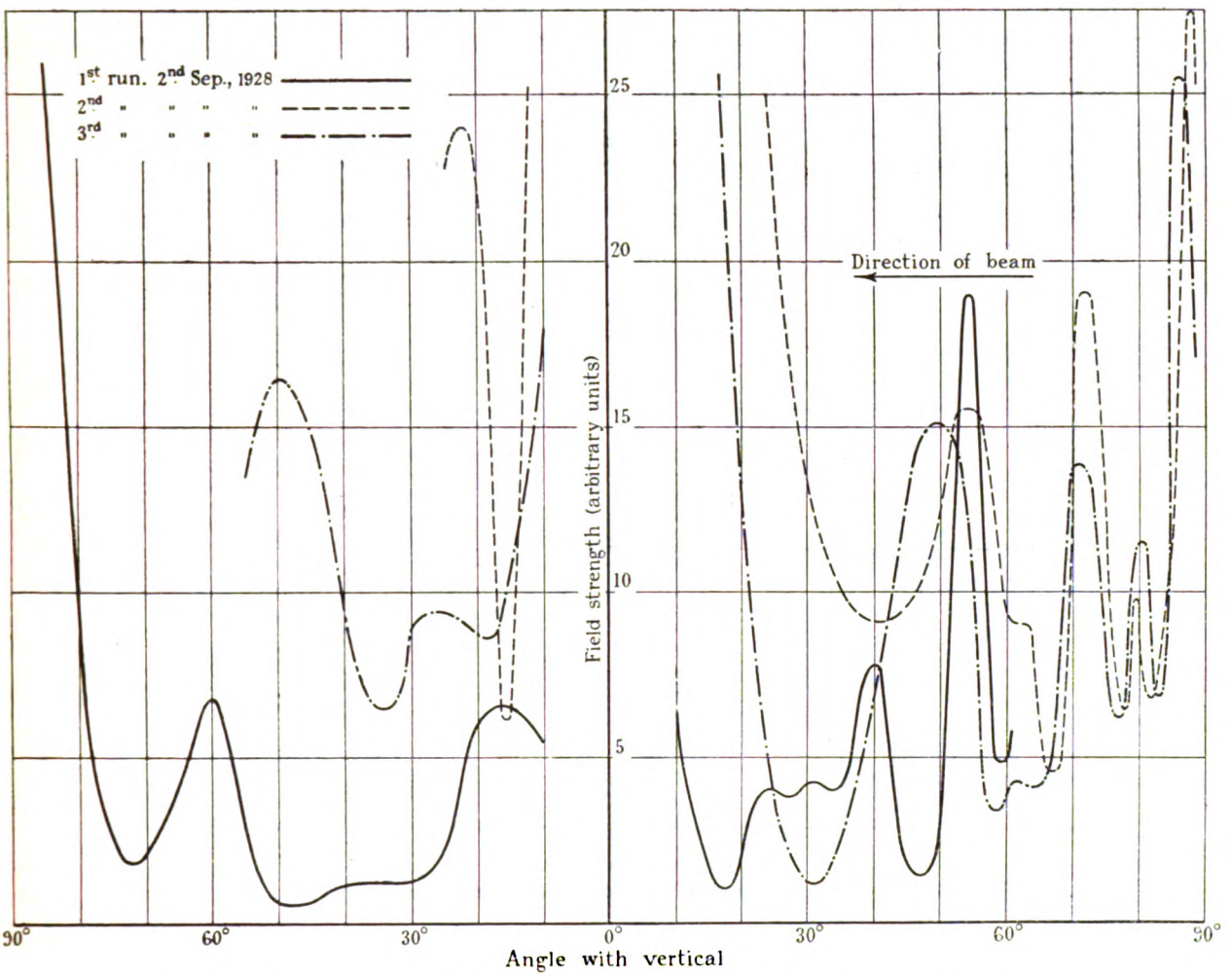

Fig. 14.—Vertical radiation distribution of Australian beam aerial at Grimsby. $\lambda = 25{\cdot}9$ m.

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vertical in the case of a perfect earth will be proportional to

$$A \cos (h_1 \cos \theta)$$

The total field strength due to the whole antenna is obtained by adding all such terms. In the case of an imperfect earth the field strength has to be calculated as explained in Appendix 2.

It should be pointed out that the radiation distribution should be identical in form but not in strength in all vertical planes passing through the centre of the antenna array, if the reflector is neglected. The reflector consists of simple straight wires 960° high situated ¼ wavelength behind the transmitting array itself. This beam is so arranged as to permit of transmission either way round the earth. It was a simple matter, therefore, to find the ratio of front-to-back radiation, by simply reversing the direction of the beam, leaving the measuring receiver in the same place. The ratio on the ground of the field strength in front and behind was found to be of the order of 2 : 1, which value roughly checked with calculations based on the current measurements made on the reflectors and the transmitters, together with a rough estimate of the phase relation between them (see Appendix 2). This showed that the reflector would not affect the radiation distribution in a vertical plane to a very large extent, and, considering the errors liable to be found, it was not considered necessary to allow for it. The experimental results are given in Fig. 14 and the theoretical curve calculated for a perfect earth is shown in Fig. 15.

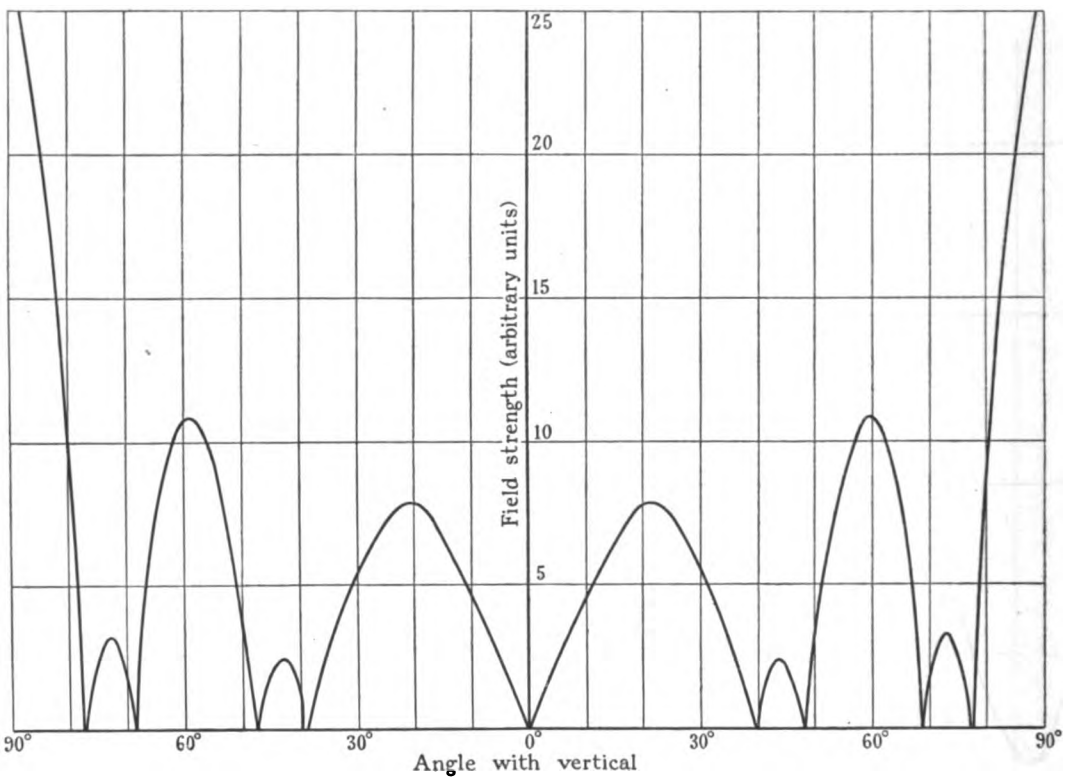

FIG. 15.—Theoretical radiation distribution from Australian beam over a perfect earth. $\lambda = 25{\cdot}9$ m.

The difficulty of investigating experimentally the radiation distribution of a beam compared with that of the simple antenna, lies in the very high directive properties of the former. Should the aeroplane fly in a plane slightly displaced from the true plane, the result would be very considerably distorted. Actually it was noted from the ground that the aeroplane flew in a plane parallel to, but slightly displaced from, the true plane. The true flight of the aeroplane was in a plane containing the main beam (i.e. perpendicular to the antenna array), and a displacement horizontally of as small an angle as 5° would reduce the signal from a strong maximum to a minimum. The size of the antenna system produced a further error, as it was not possible to reach such a height that the distance of the aeroplane from the antenna was large compared with the size of the antenna system itself. This error, however, when flying in the plane of the beam, is only of a second order. The aeroplane flight was practically a perfect straight line in space, starting at an altitude of 8 000 ft. at a horizontal distance of about 4 000 ft. behind the antenna, and ending at an altitude of some 500 ft. at a horizontal distance of about 25 000 ft. in front of the antenna. In flight 1 the signal strength was noted to be comparatively weak. This was no doubt due to the aeroplane flying appreciably out of the true plane, and it explains why the radiation distribution on this flight differs so considerably from that of the other two flights. The agreement between the other two flights is considered very satisfactory, in view of the experimental difficulties, and the results of flight 1 may, on this evidence, be discarded as untrustworthy.

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The comparison of the experimental results and theoretical curves, as shown in Figs. 14 and 15, is very interesting. The large value of field strength obtained experimentally at 0° is probably due to a small asymmetry in the excitation of the beam system, and also to the fact that the aeroplane was not flying absolutely in the true plane. The first minimum between 30° and 65° is not so good. The third at 75° no doubt corresponds to the theoretical minimum at 69°, while the fourth at 82° corresponds to that at 77°. Then follows a very interesting result. Instead of increasing to reach a maximum of field strength on the ground, the field strength reaches a maximum at 86° and falls rapidly. Such an effect would be expected if the ground were not a perfect conductor. The angular position of the maximum can be used in obtaining an estimate of the conductivity of the ground. This would appear to be of the order of 4×10^8 electrostatic units. If this figure were used for correcting the theoretical radiation distribution curve given in Fig. 15, very little difference would be found except for angles greater than 82°.

(5) Conclusion.

From these results we may conclude that the experimental values are reliable but that very accurate analysis of the differences between the experimental and theoretical results is rendered impossible by the nature of the transmitting site.

There is considerable evidence that the minor irregularities seen in Figs. 4, 6 and 8 are due to ground effects and are not to be found in the transmission from the special antenna construction used. On the other hand, it seems difficult to attribute the discrepancies between the experimental and theoretical results to the same cause, although, judging from the size of the variations noted in Fig. 10, they may well be sufficient to explain all the discrepancies. In all the cases investigated with the simple horizontal antenna, the maxima and minima were all shifted towards the vertical, and this shift increased with decrease of wavelength. This occurred irrespectively of the nature of such maxima and minima, i.e. whether they were due to the antenna alone or to the interference between the antenna and its image (as explained above).

Evidence has been given that the inverse-distance law of field strength holds for angles of elevation at least as small as $\frac{1}{2}°$.

The results on the beam antenna show comparatively good agreement with theory, considering the very much greater experimental difficulties of this case. It is interesting to note that the shift of the maxima and minima in this case is away from the vertical, which is the reverse of the results obtained on the simple antenna.

The final conclusion to be drawn from this is that the theory of radiation distribution is substantially correct, but, in order to examine its accuracy in great detail, it would be necessary to carry out experiments over some irreproachable site, and this might be more difficult to find than would at first appear. The necessity for this would in any case be somewhat academic, for in practice it might be difficult to find such a site for the erection of a transmitting station, and if this were not obtained the ordinary calculations would never give a very high order of accuracy. It would have been desirable to carry out a few more tests in order to confirm the deductions reached, but the author was unable to continue the experiments.

In conclusion the author wishes to thank the many workers who have helped him in carrying out this investigation. In particular, his thanks are due to the Air Ministry, who generously supplied the aeroplane for carrying out the tests; also to Squadron-Leader A. L. Gregory, Messrs. F. S. Barton, N. F. S. Hecht, T. H. Gill and members of their staff at the Royal Aircraft Establishment, who gave him all the facilities and advice that could be desired during the progress of the work.

The author has also to thank the General Post Office and in particular Colonels A. G. Lee and A. S. Angwin, and Mr. De Wardt, for permitting and giving every assistance in carrying out the measurement on the beam antenna.

Finally, and principally, he wishes to thank Flight-Lieut. Ragg, R.A.F., whose skill and patience in piloting the aeroplane were largely responsible for the unexpected accuracy of the measurements, and also Messrs. J. S. McPetrie, B.Sc., and B. J. Byrne, of the National Physical Laboratory, without whose efficiency and patience the work would not have reached its comparatively satisfactory conclusion.

This work was carried out as part of the programme of the Radio Research Board and is published by permission of the Department of Scientific and Industrial Research.

APPENDIX 1.

Correction for End Effects.

The author has shown experimentally* that in a straight vertical antenna the first half-loop of current was some 8 per cent greater in area than the half-loop of a sine wave. At the foot of the antenna the effect may be expected to depend on the length and the proximity to the ground. If the length is l and the wavelength λ, the charge per unit length near the top when the current at the centre of a loop is I will be proportional to I, while that at the foot will be proportional to $I \cos(2\pi l/\lambda)$.

The end effect is produced by a sudden discontinuity in the charge. At the top, from a value proportional to I it suddenly changes to zero, while at the foot it changes from $I \cos(2\pi l/\lambda)$ to $-I \cos(2\pi l/\lambda)$. When the foot of the antenna is in close proximity to the ground it is to be expected, therefore, that the end effect at the foot will be $2 \cos(2\pi l/\lambda)$ as large as that at the top. To simplify calculations we may, with sufficient accuracy, represent the end effects by an area 8 per cent as great as a half-loop of current concentrated at the top end, together with one $16 \cos(2\pi l/\lambda)$ per cent concentrated at the foot. The radiation distribution from such concentrated areas may be very readily calculated, and superposed on the radiation distribution calculated from a simple sinusoidal distribution.

The case of a horizontal antenna as shown in Fig. 3 is similar to that of a vertical antenna turned through 90°.

* See Reference (4).

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The centre point C now corresponds to the foot of the antenna. It must be remembered, however, that these end effects have their images at A′, B′ and C′ in the ground.

APPENDIX 2.

Measurement of the Phase Relations in Antenna Arrays.

For the purpose of measuring the phase difference between two antennæ a low-voltage cathode-ray oscillograph was used. E.M.F.'s from the two antennæ A and B (Fig. 16) whose phase difference is to be measured are induced in two coils C and D connected to the opposite plates of the oscillograph. An ellipse is obtained the size and ellipticity of which varies with the e.m.f.'s induced in C and D. These are adjusted until the axes

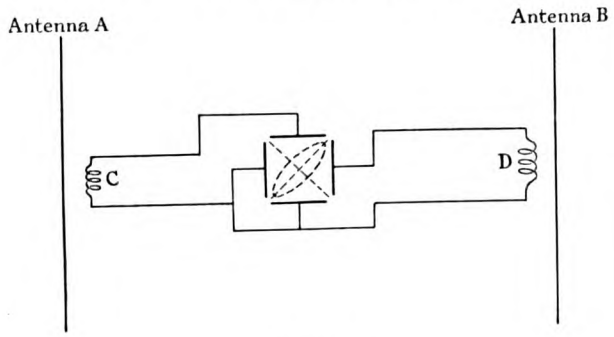

Fig. 16.

of the ellipse bisect the line joining the plates, as shown. It can readily be shown then that, if γ is the phase angle and a and b are the axes of the ellipse,

$$\tan \gamma = a/b.$$

It will be noted that an ambiguity exists, for

$$\gamma = \text{arc tan } (a/b) \quad \text{or} \quad \pi + \text{arc tan } (a/b).$$

An error may occur at very high frequencies, owing to the fact that the deflecting plates of the oscillograph are not at the same distance from the cathode. This error may be readily eliminated by interchanging the connections of the coils C and D to the oscillograph plates and taking a mean of the results.

The instrument used was placed in a metal box in which a small hole was cut for the purpose of observation. In front of the tube a number of wires were stretched at 45° to the deflecting plates. The distance between the wires was such as to show deflections in the ratio of 1: 2: 3: 4, etc., of the spot on the screen. The wires were faintly lighted by a means of a neon lamp working from the high-tension supply of the oscillograph. In this way the axes of the ellipse could be readily adjusted to the correct direction and their lengths read off. The leads and coils connected to the instrument were made as identical as possible.

Good consistency was obtained at 26 m, but considerable irregularities were found at 15 m. Such an instrument should be very valuable during the erection of a complicated beam system, but the author feels that the measurements made with it should be subjected to more thorough investigation. The phase relation between a pair of isolated, spaced antennæ can be obtained by taking measurements of the radiation distribution along the ground, and this method might be used as a check in measurements made with the cathode ray.

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SOME DEVELOPMENTS OF THE THERMIONIC VOLTMETER.*

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SUMMARY.

In this paper the various ways are considered in which a 3-electrode valve should be used in a thermionic voltmeter, and the accumulated experience is described of several years' work by the author in making valve voltmeters for various purposes. No one arrangement can possess every desirable property, and it is shown how to produce a voltmeter to fulfil specified requirements. The effective input resistance of the various arrangements, and the harmonic currents produced by grid current, are considered analytically and experimentally. The effect on range and sensitivity of providing additional batteries is also discussed, extra batteries being found to be an advantage only in very low-range instruments.

The effect on the calibration of a change in the valve and in temperature is found to depend on the form of the rectifier used. In general, a thermionic voltmeter cannot read R.M.S. values. The wave-form error of various systems is investigated; with one arrangement the reading is proportional to the peak voltage and in another to the mean. The possibility of a frequency error is discussed; tests show that the error is negligible for frequencies up to 1 000 kilocycles per sec., whilst from indirect tests it would appear that the error is probably negligible for frequencies up to 30 times this value.

(1) INTRODUCTION.

A rectifier, such as a crystal or thermionic valve, may be used to make an alternating voltage perceptible to a polarized galvanometer, and in recent years these have been combined to produce a satisfactory voltmeter. The thermionic valve, with either two, three or four electrodes, possesses great potentialities for voltmeter use, and this paper explores some of the most desirable.

The ideal a.c. thermionic voltmeter should possess the following properties. It should be as portable as a d.c. ammeter and be permanently calibrated; it should possess great sensitivity and wide range and be unaffected by a steady voltage between its terminals; its input resistance should be infinite, and its readings should be proportional to the R.M.S. value of any wave-form and unaffected by frequency.

By a suitable choice of valve, batteries and galvanometer, most of these desirable attributes can be secured, but they cannot all be secured simultaneously. Simplicity and portability are two of the most desirable attributes: the fixed location inevitable with a table full of semi-portable apparatus imposes severe limitations on the use of a thermionic voltmeter, and therefore it must be worth while to design portable instruments, even though they may possess limitations which are well understood. The same consideration underlies much of the art and skill of instrument design; for example, the silver voltameter and current balance may be the most perfect ammeters known, yet their sphere of application is vastly more limited than that of a direct-reading pivoted instrument.

We shall now consider how to make portable thermionic voltmeters having the least number of features which will in general be undesirable.

(2) ELIMINATION OF UNESSENTIAL BATTERIES AND METHOD OF REPRODUCING CALIBRATION CONDITIONS.

A thermionic voltmeter must have a battery to heat the filament and to this extent cannot be as portable as a moving-coil voltmeter. But this battery need not supply more than, say, 0·1 amp. at 3 or 4 volts; it can therefore be a small dry battery and the inconvenience of this is not overwhelming. Additional batteries are not essential to the working of the voltmeter. We shall therefore postulate a problem as follows: Given a galvanometer, thermionic valve and filament battery, to produce a voltmeter which shall possess the greatest possible number of desirable qualities.

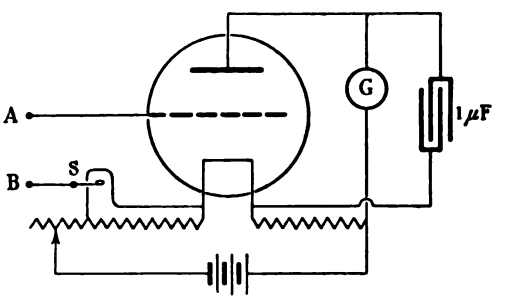

FIG. 1.

The instrument is to be permanently calibrated and provision must therefore be made to reproduce the conditions, in respect of filament current, anode and grid potential, which obtained at the time of calibration. If connections basically like those depicted in Fig. 1 are used we may reproduce the exact calibration conditions by adjusting the filament rheostat until the galvanometer in the anode circuit deflects to a marked point on the scale. It is convenient to provide a testing switch (marked S in Fig. 1) which temporarily removes the grid bias and so enhances the test-point deflection. The sensitivity of the galvanometer must at least be sufficient to produce an adequate deflection for the test point. With ordinary valves, this usually entails that the galvanometer shall be as sensitive as it is possible to make a portable unipivot instrument.

* Reprinted from *Journal I.E.E.*, 1930, vol. 68, p. 1039.

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(3) Limitations of Range.

Consider the system of Fig. 1. An alternating voltage applied between A and B produces an increase of mean anode current which is proportional to the square of the applied voltage when this voltage is very small. The lower limit of range depends, therefore, on the sensitivity of the galvanometer; in a portable voltmeter equipped with a unipivot instrument the smallest detectable current is of the order of 0·25 μA, and this will be produced by an applied p.d. of about 0·1 volt (R.M.S.) Provision of additional anode potential will not alter this sensitivity appreciably but will extend the square-law portion of the scale. Greater sensitivity can be obtained only by using several valves in parallel or by adding an amplifier. The upper limit of range is imposed by the available anode potential, and in Fig. 1

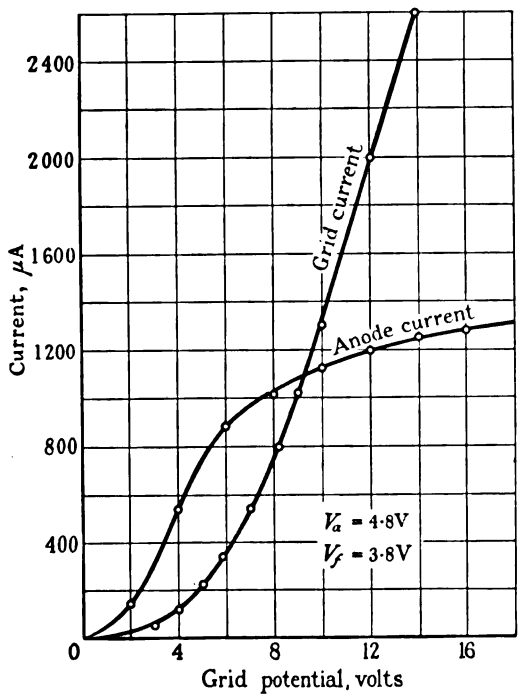

Fig. 2.

this is supposed to be 4·5 volts positive to the negative end of the filament. When the grid becomes appreciably more positive than the anode the grid current will exceed the anode current, which tends to a limiting or maximum value. This is illustrated by the curve of Fig. 2 (which relates to a P.M.4 valve connected as in Fig. 1 but with zero grid bias). Fig. 2 shows that if the applied voltage exceeds about 12 volts maximum (8·4 volts R.M.S.) the anode current increases very slowly, and it becomes necessary to move the galvanometer to the grid circuit and use the valve as a diode. Such an arrangement would be undesirable because the voltmeter would take considerable power from the source and so have a low effective resistance. Therefore the negative grid-bias must be increased, but if extra batteries are precluded this must be done by some automatic process. Two such methods are shown in Figs. 3 and 4, and of these Fig. 3 has the great advantage of isolating the voltmeter from steady voltages and so confers one of the desirable attributes postulated in the Introduction. In Fig. 3 the spurts of rectified grid current divide between the parallel paths provided by the condenser and the grid leak. The capacity of the condenser is so large that sensibly all the alternating components flow through it, leaving a substantially steady p.d. between its terminals. The combination of large capacity and high resistance is therefore precisely equivalent to a battery giving a negative bias to the grid and, moreover, a bias which increases in proportion

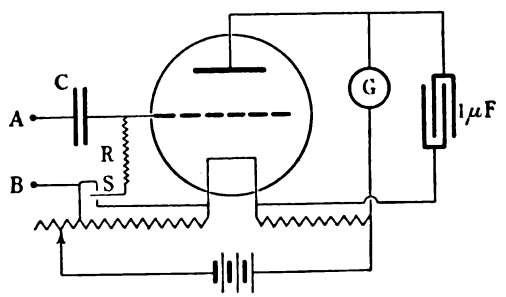

Fig. 3.

to the applied a.c. voltage. The saturation effect of Fig. 2 will now be delayed and anode current will flow during a portion only of the positive half-cycle of voltage.

In the system of Fig. 4 the positive voltage applied to the grid is but a small fraction of the voltage applied between A and B, and this fraction depends on the ratio of the effective resistance of the grid-filament path compared with that of the grid leak in series with it. Here anode current will flow during the whole of the positive half-cycle of voltage, but in effect the voltmeter is fed from a potential divider. Since no current flows during the negative half-cycle, the resistance is infinite

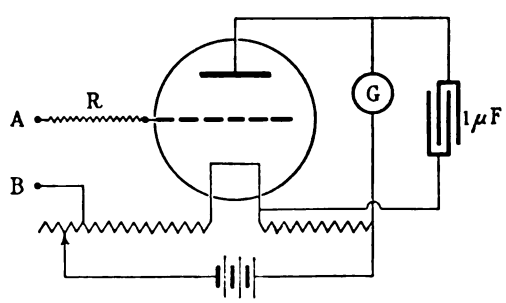

Fig. 4.

for that period and so the system is preferable to a potential divider; moreover, the variable conductivity of the grid provides a potential divider in which the tapping point is lowered automatically as the applied voltage is increased. The systems shown in Figs. 1, 3, and 4 take power from the source; we shall now consider the effective input resistance of voltmeters which use any one of these three systems.

(4) Effective Input Resistance.

The effective input resistance may become finite from any or all of four independent causes; namely, energy transferred from the grid to the anode circuit, dielectric

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losses in the valve cap and instrument generally, losses in the grid leak, and losses due to grid current flowing during a part or the whole of the positive half-cycle of applied voltage. The energy is transferred from grid to anode by means of the grid-to-anode capacity and is wasted in the resistance of the anode circuit impedance (galvanometer, etc.). The anode circuit impedance should be shunted by a large capacity, say 1 μF, and then the anode is in effect short-circuited to the filament and the anode potential does not fluctuate. The grid-anode capacity is then in parallel with the grid-filament capacity and contributes only to the input capacity without contributing anything to the input resistance. We shall presume that the anode is so connected to the filament and that then the first-mentioned cause of loss is removed.

The dielectric losses are reduced by removing the valve cap and intelligently disposing the valve and terminals in the instrument case; dielectric loss cannot be eliminated, and is troublesome to measure. The results obtained on a voltmeter constructed as in Fig. 1, but without the calibrating switch, may be quoted as an example. A circuit having an inductance of 1·14 mH was tuned to a frequency of 212 kilocycles per sec. ($\lambda = 1\,420$ m), and by plotting the resonance curve the inherent decrement was found to be 1·80 per cent. The complete voltmeter (of the type depicted in Fig. 1) was then connected across the whole of the inductance, but, to prevent grid damping, the filament was not switched on. The circuit decrement was determined again with the voltmeter connected to the circuit and was now found to be 1·85 per cent, so that the presence of the voltmeter had increased the apparent resistance of the coil by 3 per cent. But some of this 3 per cent increase is due to the capacity of the voltmeter increasing the apparent resistance of the coil in a manner analogous to self-capacity.

The two resonance curves showed the apparent input capacity of the voltmeter to be 5·5 $\mu\mu$F and it may be calculated, from the value of L and the frequency, that the tuning capacity was 485 $\mu\mu$F. Now if a condenser of capacity C_1 shunts a coil L, which is in series with a condenser of capacity C_2, and the system is tuned to resonance, it is well known that C_1 increases the apparent resistance of the coil in the ratio $[(C_1 + C_2)/C_2]^2$: in this case the ratio is $(485/479{\cdot}5)^2 = 1{\cdot}023$. Thus practically all the observed increase of decrement could be accounted for if the power factor of the voltmeter were zero. In this instrument the valve cap had been removed and the grid was connected to the input terminal by a bare lead only about 3 cm long, otherwise so small a capacity as 5·5 $\mu\mu$F would not have been achieved. As a contrast, a similar voltmeter containing a similar valve, but in which the valve cap was intact and placed in a moulded ebonite valve-holder and the grid lead was long and heavy, may be quoted. This instrument had a capacity of 19 $\mu\mu$F, and increased the apparent resistance of the same circuit by 20 per cent. It may be calculated that the shunting effect of 19 $\mu\mu$F would increase the apparent resistance of the coil by 8·2 per cent, so the residue of 11·8 per cent must be attributed to bad power factor. From the stated values of L, C, f and δ, it may be calculated that an increase of 11·8 per cent would be produced by a resistance of 2 MΩ shunting the coil. It will also be found that the impedance of 19 $\mu\mu$F at the relevant frequency is 1/25·3 MΩ, so that the power factor of the voltmeter was 0·02. If the same process is applied to the rather inadequate difference observed for the capless valve, a power factor of 0·005 will be deduced: on the other hand, if the effect of the capacity had been ignored the voltmeter would have been held to have a power factor of 0·017. So it is clear that if the cap is removed and all due precautions are taken the power factor need not exceed 0·01, and if the cap is not removed it will be about 0·02.

Concerning the power factor of valve capacities, a paper* by G. W. Sutton should be consulted. Sutton finds, for valves without their caps, power factors which range from 0·003 to 0·02, and concludes that "it should be sufficient merely to remove the cap from an ordinary receiving valve for use as a valve voltmeter." †

When the system of Fig. 3 is used, conditions are made worse both because of the conductance of the grid leak and also because of the poor power factor inevitably associated with the grid leak and condenser. As an example, a carefully constructed voltmeter arranged as in Fig. 3, but with some switching arrangements on the filament side of the leak (of resistance 1 MΩ) had an apparent capacity of 31 $\mu\mu$F and raised the decrement of the test circuit from 1·8 per cent to 3 per cent. The enhanced capacity is due partly to some switching arrangements, but is partly inevitable from the earth capacity of the rather bulky grid condenser and leak connections. It may be found that the capacity of 31 $\mu\mu$F increases the apparent resistance of the coil by 13·8 per cent and so leaves a residue of 52 per cent to attribute to the grid leak and to power factor. It may be found that this increase would be produced by a shunt resistance of 0·5 MΩ, but this is due partly to the real resistance of the 1-MΩ leak, and so the power factor is equivalent to a shunt resistance of 1 MΩ. Combining this with the impedance of 31 $\mu\mu$F at 212 kilocycles per sec., the power factor of the voltmeter is 0·024. A voltmeter arranged as in Fig. 4 and having an unshunted grid leak of 1 MΩ had an apparent capacity of 25 $\mu\mu$F, and increased the resistance of the same test circuit by 25 per cent.

In general, therefore, a voltmeter will have a power factor between, say, 0·01 and 0·03, and the capacity will approach that of the electrodes only with the arrangement of Fig. 1. Probably the power factor is substantially constant at all frequencies, so to obtain a high value for the effective resistance at radio frequencies it is important to keep the capacity small. Viewed in this way it is more important to keep the capacity small than would appear from considering the ratio of this capacity to the other capacities of the system. In default of an actual measurement of resistance, we may estimate its value by arbitrarily assigning a power factor of, say, 0·02 to the measured value of the input capacity.

Grid current must flow in the arrangement of Figs.

* "The Power Factor and Capacity of the Electrodes and Base of Triode Valves, with special reference to their use in Thermionic Voltmeters," *Proceedings of the Physical Society of London*, 1927, vol. 40, p. 14.

† See also Bibliography, (8), p. 592, Table 2.

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3 and 4 in order that their purpose may be fulfilled, and it will flow in the arrangement of Fig. 1 when the peak value of the applied voltage exceeds the grid bias of about 1·2 V.

We may deduce the wave-form of grid current for the system of Fig. 1 from a static characteristic connecting the steady voltage between the terminals of the voltmeter and the resultant grid current. Such a static curve taken from a complete voltmeter of a well-known pattern,* typified by Fig. 1, is shown in Fig. 5. From this curve we may readily obtain the wave-form of grid current which results from a sinusoidal applied voltage. Four such wave-forms are exhibited in Fig. 6, and correspond respectively to an applied p.d. of 0·71, 1, 1·42 and 2 volts (all R.M.S.). The other half of the positive half-cycle is, of course, an exact replica of that shown by Fig. 6, and the current is zero throughout the negative half-cycle. If the ordinates of the curves of Fig. 6 are multiplied by the corresponding ordinates of their appropriate voltage wave, a curve of input power may be plotted: the effective resistance can then be deduced from the mean height of this curve taken over a whole cycle of applied p.d. This has been done

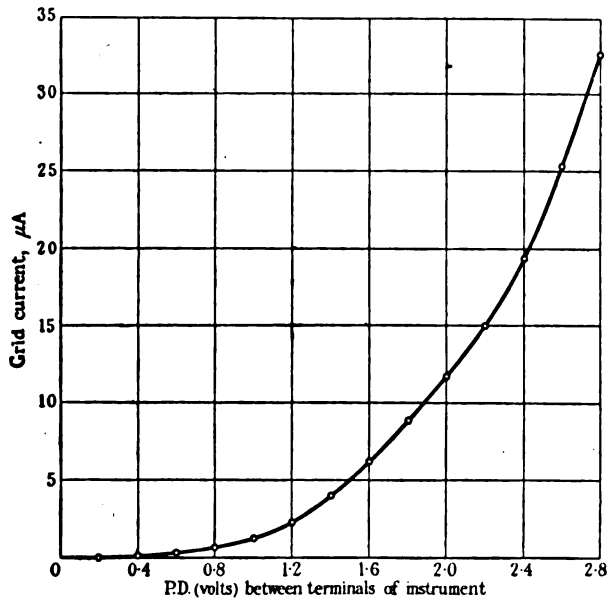

FIG. 5.

for the four curves of Fig. 6, and Fig. 7 exhibits the effective resistance plotted as a function of the R.M.S. value of the applied p.d. The full-scale deflection for this pattern of voltmeter is usually fixed at 1·5 volts (R.M.S.), and hence the effective resistance for grid damping does not fall below 0·4 MΩ.

Though it is preferable to deduce the effective resistance from curves such as Fig. 5, the resistance can be measured by observing the decrement which the voltmeter adds to a tuned circuit. As an example of such a measurement may be quoted an experience from the same measuring circuit described previously. This had an inherent decrement of 1·8 per cent at 212 kilocycles per sec. With the voltmeter across the whole of the 1·14-mH coil the decrement was increased to 3·4 per cent, and from this it follows that the effective resistance of the voltmeter was about 0·31 MΩ. This figure is in fair agreement with the deduced figure of 0·4 MΩ; and also the measurement was not made on the same instrument from which Fig. 5 was obtained. Since the resistance is a function of the applied voltage, the voltmeter will distort slightly the top of the resonance curve of the circuit to which it is attached.

The effective resistance due to power factor will vary roughly as the reciprocal of the frequency, whereas the resistance due to grid current is independent of

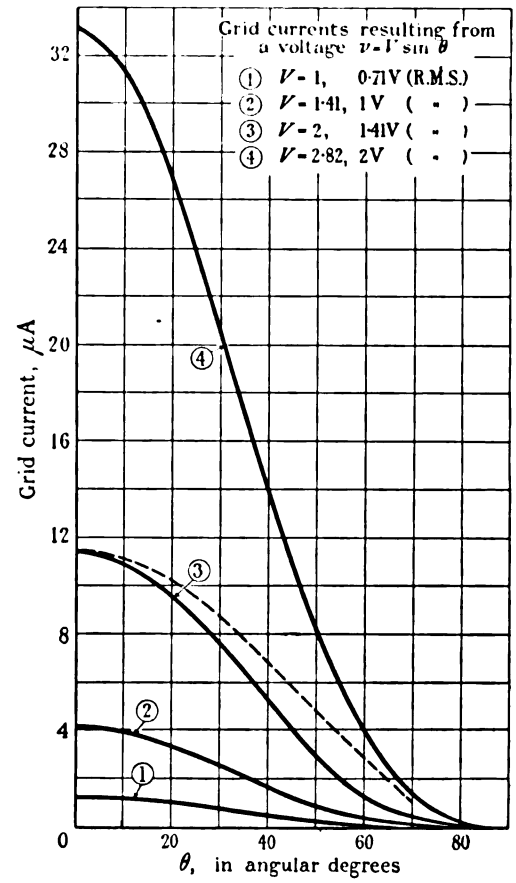

FIG. 6.

frequency. Assuming that the resistance at full scale is 0·4 MΩ for grid current and 14 MΩ for dielectric loss at 212 kilocycles per sec. (i.e. power factor of 0·01 with capacity of 5·54 μμF), we obtain the resistance-frequency values collected in Table 1.

In addition to adding effective resistance the voltmeter also creates harmonic currents and it is of some interest to estimate their value. The fundamental amplitude can be obtained at once from the deduced values of effective resistance. Thus for the four curves of Fig. 6 the maxima of the fundamental components are respectively 11·3, 4·4, 1·23, and 0·435 μA when the positive peak values were respectively 33, 11·5, 4·2 and 1·2 μA, and the ratio of positive peak to fundamental

* Moullin Voltmeter, Pattern A, No. L22474.

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amplitude does not differ much from 3. It is scarcely worth while to make a Fourier analysis of the curves of Fig. 6, but they approximate roughly to the form $y = (1 - \cos 2x)$ from $x = 0$ to $x = \pi$ and $y = 0$ from $x = \pi$ to $x = 2\pi$, and this is the equation of the dotted curve near (3) of Fig. 6. The Fourier series representing this curve is

$$y = \frac{1}{4} + \frac{4}{3\pi}\left(\sin x - \frac{3\pi}{16}\cos 2x - \frac{1}{5}\sin 3x - \frac{1}{35}\sin 5x \ldots\right)$$

According to this the second harmonic is 59 per cent and the third harmonic 20 per cent of the fundamental. If we had supposed the curves of Fig. 6 to be sine curves during the positive half-cycle and zero during the

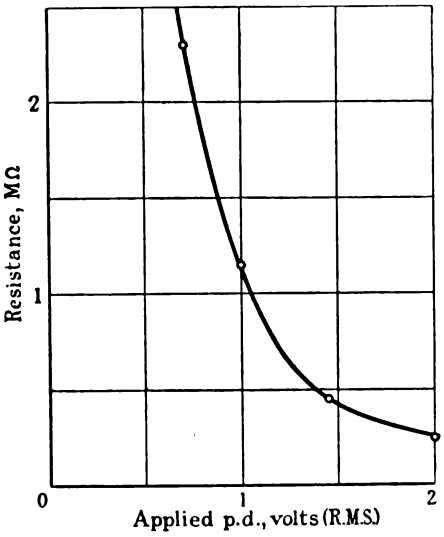

FIG. 7.

negative half-cycle, it may be shown that the second harmonic would have been 48 per cent and the third harmonic 8 per cent. Evidently the size of the second harmonic is not very sensitive to the exact shape of the curve, and we may assume roughly that the second-harmonic current through the voltmeter is about 50 per

TABLE 1.

Frequency	λ	Resistance
kilocycles per sec.	m	MΩ
1	300 000	0·4
100	3 000	0·395
1 000	300	0·376
10 000	30	0·17
30 000	10	0·08

cent of the fundamental: from this we may estimate roughly the second-harmonic current in the remainder of the circuit.

We now consider the effective resistance due to grid current in the arrangement of Fig. 3. A 6-volt filament battery will permit a negative bias of at least 1·4 volts. We shall assume that this has been provided and then there will be no grid current until the applied voltage exceeds 1·0 volt (R.M.S.). The current flowing in this condition has been studied closely by S. E. A. Landale,* who shows there that if the equation of the grid current curve is taken to be a parabola, then the mean grid

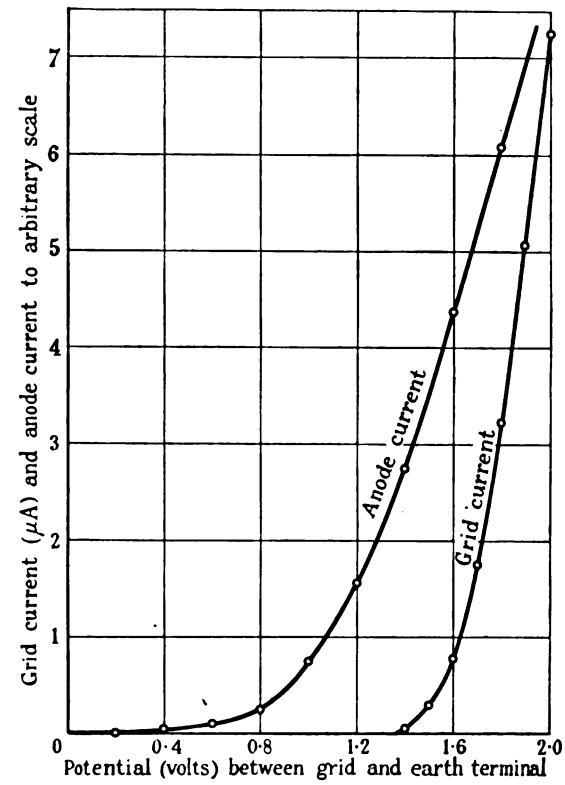

FIG. 8.

current may be predicted accurately from 0·10 volt (R.M.S.) up to any value. Reference to this paper will show how to determine the effective resistance accurately, but we shall do this roughly for the voltmeter having the static characteristic represented by Fig. 8. Over the range covered in this figure we shall assume the grid-current curve to be straight, starting at 1·4 V

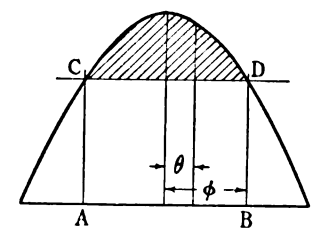

FIG. 9.

and having a slope of 15 μA/V. Let us take a grid leak of 2 MΩ and a permanent bias of − 1·4 V. Consider Fig. 9, which shows a positive half-cycle of voltage of maximum value V. Grid current flows during that portion of the cycle only which is shown shaded. Let the mean grid current be I, then grid current will begin to flow at

* See Bibliography, (15).

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the moment of the cycle when the voltage rises to a value AC given by the equation $RI + 1{\cdot}4 = V\cos\phi$. Now $I = Ab/2\pi$, where b is the slope conductance of the grid characteristic and A is the area of the shaded cap. From this we have

$$\frac{bRV}{\pi}(\sin\phi - \phi\cos\phi) + 1{\cdot}4 = V\cos\phi$$

Taking $b = 15\,\mu\text{A/V}$ and $R = 2\,\text{M}\Omega$, we have

$$\frac{30V}{\pi}(\sin\phi - \phi\cos\phi) + 1{\cdot}4 = V\cos\phi$$

On substituting in this $\phi = \pi/10$, $\pi/8$ and $\pi/6$, we find that $V = 1{\cdot}64$ V, $1{\cdot}92$ V and $3{\cdot}3$ V respectively. The instantaneous grid current is $i = V(\cos\theta - \cos\phi)b$ and the instantaneous rate of working is

$$W = iV\cos\theta$$
$$= bV^2(\cos^2\theta - \cos\theta\cos\phi)$$

$$\therefore \frac{\int_0^\phi iV\cos\theta\,.\,d\theta}{\pi} = \frac{bV^2}{2\pi}\left(\phi - \frac{\sin 2\phi}{2}\right)$$
$$= \frac{V^2}{2R}$$

where R is the effective resistance

$$\therefore R = \frac{\pi}{b(\phi - \frac{1}{2}\sin 2\phi)}$$

Giving ϕ the values $\pi/10$, $\pi/8$ and $\pi/6$ we find that R has the values 10, 5 and $2{\cdot}3\,\text{M}\Omega$ respectively. This effective resistance is in parallel with a grid leak of $2\,\text{M}\Omega$, and so the net effective values are $1{\cdot}68$, $1{\cdot}43$ and $1{\cdot}07$ $\text{M}\Omega$ respectively.

In accordance with our previous working we shall take the effective resistance due to imperfect power factor as equivalent to $1{\cdot}25\,\text{M}\Omega$ at 212 kilocycles per sec. (i.e. power factor of $0{\cdot}02$ with a capacity of $30\,\mu\mu\text{F}$) and so obtain the resistance/frequency values collected in Table 2.

TABLE 2.

Frequency	λ	Resistance (MΩ)		
		$V = 1{\cdot}64$	$V = 1{\cdot}92$	$V = 3{\cdot}3$
kilocycles per sec.	m			
1	300 000	1·68	1·43	1·07
100	3 000	0·1	0·93	0·73
1 000	300	0·22	0·22	0·20
10 000	30	0·026	0·026	0·026

Comparing this with Table 1 we see that the grid-leak system of Fig. 3 is better than the arrangement of Fig. 1, except for frequencies which make the effect of imperfect power factor all-important. This is a point worth careful notice and is of general application in radio work. Since we find ϕ is a small angle, we can approach the matter more simply and in a way which leads to a simple experimental method of determining the effective resistance. Since grid current does not start to flow until the voltage has reached the value $V\cos\phi$, and since $\cos\phi$ is very nearly unity, we can write

$$\frac{\int_0^\phi iV\cos\theta\,.\,d\theta}{\pi} \fallingdotseq \frac{V\int_0^\phi i\,.\,d\theta}{\pi} = VI$$

where I is the mean grid current.

$$\therefore VI = V^2/2R$$

whence $$R = \frac{V}{2I} = \frac{V_{R.M.S.}}{\sqrt{2}I}$$

If, therefore, a curve connecting rectified grid current and R.M.S. voltage be plotted from experiment the effective input resistance is found by dividing the voltage by $\sqrt{2}$ times the corresponding mean grid current. For large applied voltages, say $V > 14$, the curve of grid current and voltage becomes a straight line which just does not pass through the origin, and then I approaches very closely to the value $I = V/R$, where R is the grid-leak resistance. This is very nearly true for all valves and all reasonable values of R.*

Then we have $$R \fallingdotseq V/(2I)$$
$$\fallingdotseq \tfrac{1}{2}R$$

and so the effective resistance tends to a limiting value equal to half that of the grid leak. This simple formula must always under-estimate the effective resistance, because I is always a little less than V/R and not all the grid current flows at quite the full voltage V. Hence the effective resistance must be more than half that of the grid leak. Since this effective resistance is in parallel with the grid leak the net effective resistance must always be greater than $\frac{1}{3}R$. If, therefore, the voltmeter has a 3-MΩ grid leak the effective resistance cannot fall below $1\,\text{M}\Omega$ and must always be less than $3\,\text{M}\Omega$. In parallel with this there is the effective resistance due to imperfect power factor, and this is a function of frequency, probably varying inversely as the frequency. S. E. A. Landale calculates (*loc. cit.*, Table 3) the effective resistance due to grid current for a certain D.E.L. 410 valve combined with a 1-MΩ grid leak and finds the values shown collected in Table 3.

TABLE 3.

Applied p.d.	Effective resistance
volts (R.M.S.)	MΩ
63·1	0·57
20·6	0·58
7·74	0·62
1·77	0·72
0·24	1·75

Table 3 shows that the limiting value of $\frac{1}{2}R$ is approached very slowly and that it is more appropriate to take a figure of about $0{\cdot}65$ than one of $0{\cdot}5$.

* Bibliography, (11), p. 887; also Bibliography, (15).

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The oscillograms shown in Figs. 10 (*a*) and 10 (*b*) are instructive in illustrating the foregoing analysis. Fig. 10 (*a*) shows the current flowing through the grid leak for an applied voltage of 12 volts (R.M.S.) and with a permanent bias of $-1{\cdot}0$ V. The top edge of the thick black line is the zero of the record and the oscillograph spot would lie on this when no voltage was applied. It will be seen that the current flowing is sensibly a sine curve superposed on a steady current. A thin line has been drawn half-way between the tops and bottoms of the sine curve, and the distance between this line and the top of the black zero line represents the mean rectified current which we have called I. Measurement on the oscillograms shows that I is 86 per cent of the maximum superposed alternating current, and this illustrates the truth of our approximate equation $I = V/R$. Grid

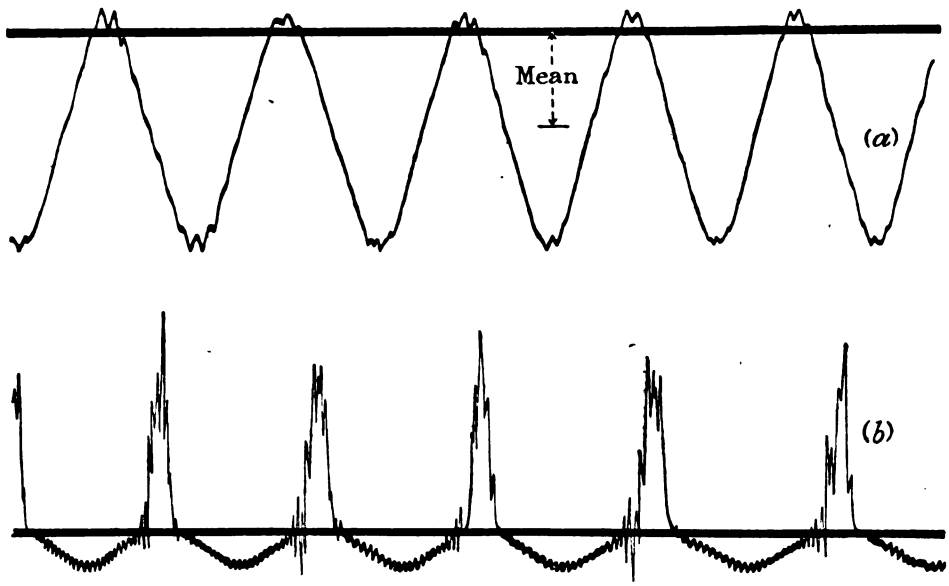

FIG. 10.

current would flow during those portions of the sine curve which lie above the zero line. Fig. 10 (*b*) is an oscillogram of the current entering the condenser from the source, that is, the input current to the voltmeter, and the mean value of this current is zero. If the grid current were zero this would be a sine curve equably disposed above and below the zero line, and if a grid-bias battery of voltage RI had been placed between the grid leak and the filament the curve would still have been a sine curve disposed equably above and below a zero line depressed an amount proportional to RI. In reality, spurts of grid current flow and are shown in Fig. 10 (*b*) by the high peaks above the zero line; the duration of these peaks corresponds to the duration of the shaded cap in Fig. 9. To obtain the requisite sensitivity (the currents being but a few microamperes) it was necessary to use an Einthoven galvanometer of relatively low natural frequency; consequently the wave-form of the peaks is not faithful and is distorted by the oscillograph.* Rough measurements from Fig. 10 (*b*) show that the grid current lasted for about one-sixth of the whole cycle, and this corresponds to $\phi = 30°$ in Fig. 9.

To make a rough estimate of the harmonic currents produced by the voltmeter, we shall assume the wave-form of the current passing from the grid into the filament to be that shown in Fig. 11. The Fourier series which represents this is

$$y = \frac{a\pi}{144}\{1 + 3{\cdot}25 \sin x + 13 \cos 2x - 13{\cdot}3 \sin 3x + 6{\cdot}5 \cos 4x - 3 \sin 5x \ldots\}$$

where, of course, $a\pi/144 = I \fallingdotseq V/R$.

According to this, at the moment of maximum voltage there is a fundamental component of current equal to $3{\cdot}25\ V/R$ entering the grid and passing to the filament, and a fundamental component of current equal to V/R entering the grid leak and passing to the filament. The fundamental component entering the condenser is therefore equal to $4{\cdot}25\ V/R$ and this leads to an effective resistance of $0{\cdot}236\ R$. This value is less than our lower limit of $\frac{1}{3}\ R$, partly because, in reality, ϕ is much less than 30°. According to the assumed wave-form the second harmonic will be $13/4{\cdot}25 = 3$ times the fundamental, the third harmonic will have about the same value and the fourth harmonic will be about $1\frac{1}{2}$ times the fundamental. We may therefore regard the voltmeter as being roughly equivalent to a resistance $\frac{1}{3}\ R$ and taking a second- and third-harmonic current of value about $9\ V/R$.

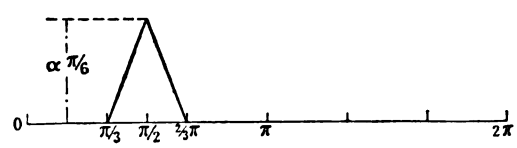

FIG. 11.

* The author is indebted for these oscillograms to the courtesy of the research department of Messrs. Cambridge Instrument Co., Ltd., London and Cambridge.

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We shall now consider the grid current taken by the arrangement of Fig. 4, and this system presents much less difficulty than either of the other two. If the voltage is applied to the grid in Fig. 4 through a resistance of the order of 1 megohm, the valve itself adds to this resistance a negligible amount when the grid is positive and an infinite amount when the grid is negative. This is capable of very definite experimental proof as follows: Let a curve be plotted connecting the grid current with the steady voltage applied between the terminals A B. The result is a straight line which misses the origin by only a small fraction of a volt.* The slope of this line will differ inappreciably from the resistance of the grid leak. (In the example quoted above, the voltage-drop across the valve was in general less than 2 per cent of the applied p.d. when the grid leak had a resistance of 1·65 MΩ.) Consequently the grid current will be $i = (V/R) \sin pt$ during the positive half-cycle and zero during the negative half-cycle. The static curve is, of course, sufficient to substantiate this, but a further check can be obtained from an a.c. measurement as follows. The maximum grid current will have the value V/R and the mean during a half-cycle will be $2/\pi$ of this, and so the mean during a whole cycle will be $I = V/(\pi R)$. An example of such a test, in which R was 1·65 MΩ, is shown by the figures given in Table 4.

The oscillogram of Fig. 12 shows the wave-form of grid current for 100 volts (R.M.S.), and the half-sine curves followed by blank spaces are apparent. The dark line which falls below the zero is due to the overshoot of the underdamped Einthoven fibre.

We obtain the effective resistance $R_{eff.}$ as follows.

$$\text{Mean power} = \frac{\int_0^{\pi} (V^2/R) \sin^2 x \, dx}{2\pi}$$

$$= \frac{V^2}{4R} = \frac{V^2_{\text{R.M.S.}}}{2R}$$

$$\therefore \quad R_{eff.} = 2R$$

The harmonic analysis of the grid current is here very definite and is

$$i = \frac{V}{\pi R}\left[1 + \frac{\pi}{2}\sin pt + \frac{1}{8}\sin 3pt + \frac{1}{24}\sin 5pt + \frac{\sqrt{5}}{3}\sin(2pt + \lambda_2) + \frac{\sqrt{5}}{15}\sin(4pt + \lambda_4)\right]$$

* For an example of such a curve see Bibliography, (11), Fig. 8.

The voltmeter therefore behaves as a constant resistance of value $2R$ and passes a second-harmonic current of amplitude $0{\cdot}235\ V/R$ and higher harmonics of negligible amplitude.

(5) Use of the Anode Current and Considerations when to Use Extra Batteries.

Grid current flows in each of the three systems considered, and in that of Figs. 3 and 4 it is essential for the purpose of extending the range beyond that suitable for Fig. 1. If, therefore, there is a rectified grid current, should it not be used to actuate the indicating galvanometer? This turns very largely on the sensitivity of the galvanometer to be used as the indicator, and also partly on the desired scale shape. The most sensitive portable galvanometers require about 20 μA for full-scale deflection, and the a.c. voltage required to give this rectified current will dictate the range of the most sensitive voltmeter we can make. Referring to Fig. 5, we may estimate that 2 volts (R.M.S.) would produce a mean grid current of about 8 μA, and 1·4 volts (R.M.S.) a mean grid current of about 2·5 μA. This system is unsuitable for voltages above 1·5 volts (R.M.S.) because the input resistance then becomes inconveniently small, so that if the grid current is indicated we could not get more than about 15 per cent of the available deflection. The rectified anode current is, however, much larger than the rectified grid current,

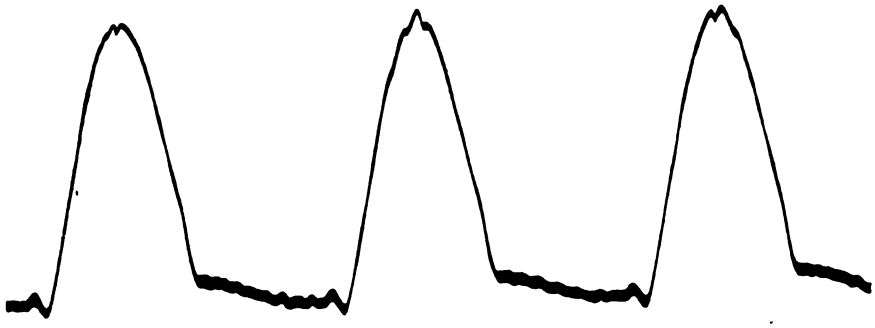

Fig. 12.

Table 4.

Applied p.d.	Measured value of I	Deduced value of R
volts (R.M.S.)	μA	ohms
240	65	1·66
200	55	1·64
180	48	1·68
80	21	1·70
46	12	1·71

and for the bright-filament valve to which Fig. 4 refers full-scale deflection may be obtained for 1·5 volts (R.M.S.). In this case, therefore, it is essential to use the anode current. Conditions would not be much improved by using an anode battery of considerable voltage. If this were provided, the anode current with no applied a.c. voltage would be considerable and a balanced

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system would have to be used. The rectified anode current (μA) would be given approximately by the equation $I = 16\ V^2$* and so there would be about 35 μA for 1·5 volts (R.M.S.) applied. We could then obtain full-scale deflection for 1·1 volts instead of 1·5 volts, but the scale would be of the square-law type, which in general is inconvenient. There would be but a trifling increase of sensitivity and considerable loss of portability and convenience.

Let us turn now to the arrangement of Fig. 3. Since the negative bias automatically increases with the applied voltage, it will be still more necessary to use the anode current with Fig. 3 than with Fig. 1. Since the bias grows progressively with the applied voltage, ultimately there will result an anode current which flows during only a portion of the positive half-cycle, this portion being slightly wider than the shaded cap of Fig. 9. This is illustrated by the oscillogram of Fig. 13, which shows the spurts of anode current with an applied p.d. of 12 volts. The combination of this effect with the effect illustrated by Fig. 2 causes the rectified anode current to reach a limiting value for an applied voltage of about 70 volts (R.M.S.). But for voltages of the order of 70, the rectified grid current will be ample to indicate directly since it is sensibly equal to the maximum of the applied voltage divided by the grid-leak resistance.

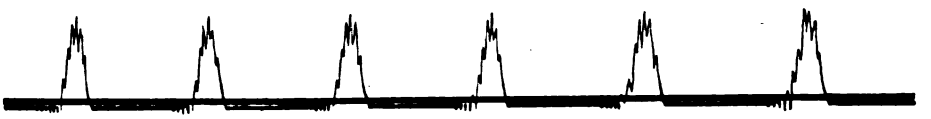

FIG. 13.

Thus with a 3-MΩ grid leak (giving the voltmeter an effective resistance of about 1 MΩ) full-scale deflection of 20 μA will result from an applied p.d. of 42 volts. The scale will be accurately linear and will be useful for voltages above about 5. If we are content with a leak of 1 MΩ (effective resistance $\frac{1}{3}$ MΩ) we can obtain a full-scale deflection for 14 volts and the scale will be useful for voltages greater than about 1·5. The system of Fig. 3, but with the galvanometer in series with the grid leak, therefore becomes naturally the next step of the range to that of Fig. 1. The decision of when to use grid current in preference to the anode current is associated with the desired input resistance, but it is probably correct to say that the grid current should be used when measuring voltages greater than about 5. Using dull-filament valves in the arrangement of Fig. 3, it is convenient to make a two-range voltmeter having a range of 0–5 volts with the galvanometer in the anode circuit, and a range of about 0–30 with the galvanometer in the grid circuit.

Is it any advantage to provide additional batteries for the system of Fig. 3? If the voltmeter is to be unaffected by steady voltages—and this was one of the desirable attributes postulated—the input system of Fig. 3 must be used. To obtain the greatest possible input resistance, grid current must be prevented by adding a grid bias battery of voltage exceeding the peak value of that applied. This battery will increase the input resistance from $\frac{1}{3}R$ to R for acoustic frequencies, but for very high frequencies the effect of imperfect power factor will swamp the advantage gained. We shall, therefore, surely decide to abandon a separate grid-bias battery. If grid current is permitted, the decrease of mean grid potential is roughly equal to the peak value of the applied voltage, and this is substantially true even down to 0·3 volt (R.M.S.).* With an anode battery of adequate voltage there need be no rectification of anode current and the decrease of mean anode current will be g times the decrease of grid potential, where g is the mutual conductance of the valve. Taking g as 2 mA/V it should be possible to get a decrease of anode current of about 2·8 mA for an applied voltage of 1 volt (R.M.S.) and the scale will be approximately linear. This is in striking contrast with the previous best figure of 16 μA for 1 volt (R.M.S.) with a square-law scale. The initial anode current will, however, now be several milliamperes and a balanced system is imperative. Also this arrangement is not capable of an extensive range, for the linear portion of the anode-current characteristic will soon be exceeded and the mean grid potential will correspond to zero anode current. Rectification of anode current will counteract the decrease of mean grid potential and there will be some applied voltage for which the indicator readings will commence to reverse. This condition can be delayed by an increase of anode battery, but it cannot thereby be postponed indefinitely.

So it seems that additional batteries are advantageous and worth while only for voltmeters in which full-scale deflection is to be produced by a small fraction of a volt. If a full-scale deflection for 1·5 volts (R.M.S.) is sufficient sensitivity, then it is both unnecessary and very inconvenient to use any batteries other than the 6-volt accumulator for the filament. Coming now to the arrangement of Fig. 4, the effective resistance of which is 2 R and in which $I = (\sqrt{2})V/(\pi R)$, it will be sufficient if R is 1 MΩ and then we shall have $I = 20\ \mu$A for $V = 45$ volts, and the grid current should be indicated for voltages greater than about 5. There are some advantages, in respect of scale shape, in indicating the anode rather than the grid current; these are discussed in the appropriate section. If a high-voltage anode battery is used with this system, there would result sinusoidal decreases of anode current and a relatively small increase having a complex wave-form. The mean decrease of anode current would be given approximately by $I = (\sqrt{2})Vg$. This relation would hold over a limited range only. For very small applied voltages the decrease would not vastly dominate the increase of anode current, and for values of V of 1 or 2 volts the anode current would be decreased to zero before V had reached its negative maximum. The scale would there-

* See MOULLIN: "Radio-Frequency Measurements," chap. 2, p. 32.

* See Bibliography, (15).

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fore be linear over a limited range, and in this range its slope could be about 1 mA per volt applied. The arrangement would require a balanced system but might be useful in very special circumstances. As a general purpose voltmeter there is nothing to be gained by using an anode battery with the system of Fig. 4.

(6) Frequency Error, Permanence of Calibration and Effect of Temperature.

The indications of the system of Fig. 1 are independent of frequency up to a very high value, because the time of flight of an electron from the filament to the grid or anode is short compared with a micro-second. The lower limit of frequency is imposed by the periodic time of the galvanometer movement. If, therefore, the voltmeter is calibrated at, say, 50 cycles per sec. the calibration probably remains unaltered up to many

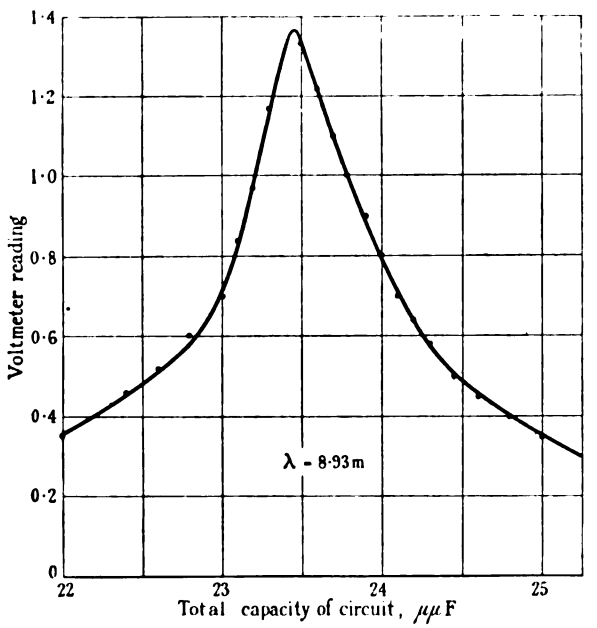

Fig. 14.

millions of cycles per second. It is not easy to make a direct test of the validity of this statement, both because electrometers of like sensitivity are not readily arranged and because their readings are not necessarily correct at the upper end of the present spectrum of frequency. A direct comparison has been made with an electrometer for frequencies up to 1 megacycle per sec.* The accuracy of comparison was such that a disagreement of 1·5 per cent could have been detected and this was not observable at any frequency up to 10^6 cycles per sec. Fig. 14 shows a resonance curve taken at a frequency of $30{\cdot}5 \times 10^6$ cycles per sec. ($\lambda = 9{\cdot}83$ m). The ordinates are the readings of a voltmeter coupled loosely to the measuring circuit, and the abscissæ are the values of the total circuit capacity. Inspection of the shape of the resonance curve shows that relative calibration has remained correct up to frequencies of over 30 megacycles per sec. Since the absolute value of the calibration has remained correct over a change of frequency from 50 to 10^6 cycles per sec., it has probably remained absolutely correct in the change from 10^6 to 30×10^6 cycles per sec., since we have experimental proof that the relative calibration has remained unchanged. We may regard such a voltmeter as the standard at such frequencies, for the capacity of an electrometer would often prohibit its use and the inductance of its suspension would probably vitiate its readings.

If these voltmeters are used at frequencies of the order of 30 megacycles per sec. a deflection may be found when the terminals are short-circuited. Apparently this is due to a voltage induced in the grid-bias coils, and the deflection is removed by shunting them by a suitable capacity. Voltmeters so provided may be used with as little care at 30 megacycles per sec. as at 100 kilocycles per sec.

The calibration of the system of Fig. 3 is also independent of frequency: so far as the response of grid or anode current to change of grid potential is concerned this follows from the tests of the system of Fig. 1. We have then to consider only whether the a.c. voltage applied to the grid remains the same fraction of that applied between A and B at all frequencies. This condition is fulfilled by making the impedance of the grid condenser sufficiently small at the lowest frequency to be used: as the voltmeter and grid leak in parallel have an effective resistance $\frac{1}{3} R$ we must so arrange C that $\frac{1}{3} R >> 1/(pC)$. If we make $\frac{1}{3} R = 5/(pC)$ at, say, 50 cycles per sec., then the calibration made at 50 cycles will be 2 per cent low at an infinite frequency. If $R = 3$ MΩ this condition is fulfilled when $C = 0{\cdot}016$ μF. A voltmeter of the pattern typified by Fig. 1 was connected in parallel with one typified by Fig. 3 and the readings of the two instruments were compared for different frequencies up to 6×10^6 cycles per sec. No difference was detectable at a deflection of 1·5 volts. In comparing two voltmeters connected in parallel at frequencies of the order of 6 megacycles per sec., due attention must be paid to the inductance of the lead to the grid-leak instrument, for this may easily cause the p.d. applied to the grid to differ from that applied to the other instrument, even though the two are separated by but a few inches.

The distributed capacity of the resistance R makes the calibration of instruments of the pattern typified by Fig. 4 dependent on frequency. At very high frequencies this distributed capacity will be sufficient to convert Fig. 4 effectively into Fig. 3 and thus increase the mean grid current from $I = (\sqrt{2})V/(\pi R)$ to $I = (\sqrt{2})V/R$, and a direct test shows that the factor π has sensibly been reached at a frequency of 2 megacycles per sec. There will be an intermediate range of frequency where the correction factor may behave in a complicated manner; however, this pattern of voltmeter is not intended for use at frequencies above a few kilocycles per second.

The permanence of calibration depends on the permanence of the valve characteristic and also on that of the grid leak (when there is one). The patterns of voltmeter typified by Figs. 3 and 4 are almost independent of the valve if this is used as a diode: this follows because $I = (\sqrt{2})V/R$ and $I = (\sqrt{2})V/(\pi R)$ respectively. Also it is capable of simple experimental proof, for it will be found that the calibration is independent of filament

* See E. B. Moullin: "Radio-Frequency Measurements," chap. 2, p. 38, Table 3.

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current over an enormous range: the range is so great that it is superfluous to provide a test-point condition and it is sufficient to use an accumulator of the correct nominal voltage. For these patterns, ageing of the valves need not be considered, but the permanence of calibration is directly proportional to the permanence of the resistance of the grid leak. Modern grid leaks are very permanent in value but have an appreciable temperature coefficient, and for this reason it is difficult to guarantee an absolute accuracy of closer than ± 2 per cent. It should, however, be noticed that even large changes of grid-leak resistance will not alter the relative calibration of the scale.

The permanence of calibration of the pattern typified by Fig. 1 does depend on the permanence of the valve characteristic, but experience shows that the calibration remains permanent for many years, provided the valves are aged for a few hours before calibration and are run with less than the rated filament current. A certain voltmeter tested by the author seven years after calibration was found to be unchanged between 1·5 V and 0·4 V, but below this the calibration had changed slightly. Possibly the satisfactory permanence of calibration is due partly to the method of adjusting the instrument zero by bringing the anode current to a predecided initial value: perhaps the filament current necessary to produce this increases by a very small amount as the valve gets older, but if this be so there is evidently no appreciable change of shape of the characteristic. When the anode current is indicated in the systems of Figs. 3 and 4, then calibration will depend on the permanence of the valve.

The effect of temperature has been studied by enclosing the voltmeter in a large case in which the air could be raised to various temperatures between + 8° C, and + 40° C. A given air temperature was maintained for a considerable period of time and it was then assumed that the whole voltmeter had reached the air temperature. An instrument typified by Fig. 3 and used as a diode was found to have a mean temperature coefficient, between 14·5° C. and 40° C., of 0·23 per cent per deg. C. An instrument typified by Fig. 4 and used as a triode was found to have a mean temperature coefficient, between 8° C. and 40° C., of 0·15 per cent per deg. C. A temperature coefficient of about + 0·2 per cent per deg. C. is probably about an average value for grid-leak voltmeters, and therefore their absolute accuracy should be reliable to ± 2 per cent if they are not exposed to a temperature range of more than ± 10 deg. C. from 15° C. Voltmeters typified by Fig. 1 have no measurable temperature coefficient.

(7) Dependence of Calibration on Wave-Form.

It is usual for an a.c. voltmeter to read the R.M.S. value of a wave of any shape, but a thermionic voltmeter does not generally do this. If the equation relating the instantaneous current i through the valve with the instantaneous value v of the applied voltage is $i = bv + cv^2$, then the mean current $I = cV^2$, where V is the R.M.S. value of the applied voltage. Valve characteristics approximate to a parabola over a very limited range, and so at best the voltmeter reading will be only roughly proportional to the R.M.S. applied voltage. If the system of Fig. 1 is provided with a suitable anode battery the indicator readings will be proportional to the R.M.S. voltage if the peak value does not exceed 2 or 3 volts,* but with ordinary valves it is difficult to extend this range very much. It must be recognized that ordinary thermionic voltmeters do not measure the R.M.S. value and it is merely fortuitous good fortune if their readings are approximately proportional thereto. It so happens that the indications of the system of Fig. 1, when employed with bright-filament valves, are not very sensitive to wave-form so long as the wave-form is not extraordinarily non-sinusoidal, for example, not more so than that ordinarily obtained from dynamos or from the magnetizing current of saturated iron. A test of this is recorded† on a wave having an in-phase third harmonic of about 30 per cent, and then the true R.M.S. value was about 1 per cent greater than that recorded by the voltmeter. A test of the curve connecting anode current and applied d.c. voltage shows that the equation is sensibly parabolic over the range ± 1 V, and so the convenient property of this pattern is explained. If the applied wave contains even harmonics the positive and negative peaks are likely to have different values, and then a reversal of connections to the voltmeter will generally yield different readings and little useful information can be deduced from either. As an example, a voltmeter of the type of Fig. 1 was connected across a resistance to measure the input current to a voltmeter of the type of Fig. 3. With the valve in type 3 switched off, the voltmeter reading was 0·52 V, from which it was deduced that the grid-leak resistance was 1·64 MΩ (the value measured by a bridge was 1·63 MΩ). With the valve switched on, this reading became 1·4 V or 1·62 V, according to the sense of connection. Reference to the oscillogram, Fig. 10 (*b*), will at once explain the difference of readings but does not suggest how to interpret them.

If the anode current is indicated in the system of Fig. 3, the reading must be very sensitive to wave-form and quite uninterpretable in terms of R.M.S. value. The anode current flows only during portions of a cycle (see oscillogram, Fig. 13), and the indications will be some function of the shape of the crest of the wave and quite independent of the shape of its main portion.

The mean anode current in the arrangement of Fig. 4 will also be dependent on wave-form, but is not nearly so dependent as in Fig. 3 because it will flow during the whole of the positive half-cycle of applied voltage. Experiment shows that the calibration is roughly independent of wave-form for such wave-forms as are ordinarily produced by dynamos. Thus a wave-form having an 8 per cent fifth harmonic‡ produced a deflection 1·8 per cent less than was produced by a sinusoidal voltage of the same R.M.S. value. But a wave-form having an anti-phase third harmonic of 34 per cent and an anti-phase fifth harmonic of 8 per cent § produced a deflection too small by 5 per cent. The shape of the scale makes the fractional change of voltage indicated greater than the fractional change of deflection. Experiment shows

* See Moullin: "Radio-Frequency Measurements," chap. 2, p. 32, Fig. 21, and G. W. Sutton: *loc. cit.*, Fig. 3.
† See Moullin: *Ibid.*, p. 36, Fig. 29.
‡ See Moullin: *Journal I.E.E.*, 1928, vol. 66, p. 891, Fig. 11B.
§ See Moullin: *loc. cit.*, Fig. 11A.

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that the relative values of the calibration are unaffected by wave-form.

When, however, the systems of Figs. 3 and 4 are used as diodes, such voltmeters measure specified properties of the wave-form and yield more information than if they measured the R.M.S. value. The relation $I = \sqrt{2}V/R$ for Fig. 3 shows that the mean grid current is proportional to the peak value of a sine curve, and it is evident from a consideration of Fig. 8 that it will be roughly proportional to the peak value of any wave-form. This has been examined experimentally in a paper by the author,* where it is shown that the readings of several very complex wave-forms are proportional to the peak value to an accuracy of better than $\pm$ 3 per cent. These experiments are supported by a rough analysis, but a more exact analysis† shows that the voltmeter must measure the peak value with an accuracy even greater than that deduced by experimental methods.

The system of Fig. 4 necessarily measures the mean value of a half wave, and this follows from the linear relation between grid current and applied d.c. voltage. The only cause of error would be from a higher harmonic of frequency sufficient to convert Fig. 4 effectively into Fig. 3, but as a given harmonic contributes to the mean in the inverse ratio of its order number the magnitude of this possible error is very small.

Sometimes the inability to read R.M.S. values with a thermionic voltmeter is annoying, and sometimes it is helpful. The times when it is annoying are very few. Low-range voltmeters are required mostly at radio frequencies where the available power and permissible capacity are small; at radio frequencies most measurements are made in resonant circuits having a low decrement and these necessarily filter away harmonics to very minute values. If a high-range voltmeter (say for more than 40 V) is being used at acoustic frequencies, presumably it is preferred to an electrometer because it has the power of indicating distortion of wave-form. To sum up, therefore, we may say that "wave-form error" is very seldom troublesome or deceptive so long as thermionic voltmeters are operated by those who understand their properties and characteristics.

(8) The Scale Shape and Multi-Ranging.

The most desirable shape of scale depends on the purpose for which the voltmeter is required. If it is used mainly as an indicator of a maximum, then a square law is convenient, but the useful range is very limited. To obtain a wide range on one scale it is preferable for the scale to be open at the start and to close up towards the top. The information contained in this paper should suffice to show how to deduce from the static characteristic the scale shape for any type of thermionic voltmeter. The scale shape for voltmeters of the type of Fig. 1 is roughly parabolic, but the openness of the upper part of the scale can be curbed by inserting a high resistance in the anode circuit; by this means about 50 per cent of the scale can be made linear. The same result can be achieved more elegantly and usefully by the reflex system of Medlam and Oschwald,* illustrated diagrammatically in Fig. 15. Here the anode-circuit resistance R not only lowers the anode potential but also provides an increasing negative bias to the grid. It should be noticed that this system cannot be used without a separate anode battery, and it will be subject to wave-form errors roughly corresponding to Fig. 3 used as a triode. When the systems of Figs. 3 and 4 are used as triodes the scale closes towards the top. The range 0–60 V is readily covered on two scales: the low range is useful from 1 to 4 V, and the second scale has an overlap on the first of 15 per cent of the maximum deflection. This arrangement is undesirable if wave-form error is important, and then it is best to achieve the second scale by using the valve as a diode. The first scale then remains 1–4 V and the second scale is linear from 0 to 16 V, with an overlap of the first of about 10 per cent of full deflection.

The system of Fig. 4, when used as a triode, is very convenient for obtaining an enormous range by simply shunting the galvanometer. The range 0–200 V may be covered in three stages, viz. 0–4–20–200 V; the

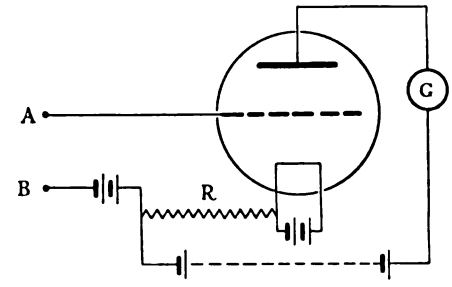

Fig. 15.

overlap between the first and second range is 35 per cent of full deflection and between the second and third is 25 per cent. Width of range in three stages is the only reason for using the arrangement of Fig. 4 as a triode. When the arrangements of Figs. 3 and 4 are used as diodes the calibration is linear.

The convenience of a voltmeter is enhanced enormously if the scale is engraved directly and so obviates an appeal to a calibration chart. It is undesirable to have more than two scales on one dial, and it is desirable that a change of range can be charted by a constant multiplying factor over the whole of at least one scale. It can be shown that this condition is fulfilled only if the equation connecting deflection θ with applied voltage V is of the form $\theta = AV^n$, where the index n has any real value. It is therefore possible to test whether one scale can be used for two ranges by making a logarithmic plot of θ and V and choosing that portion, if any, which becomes a straight line. It happens that the arrangement of Fig. 4 used as a triode does conform very closely to this condition over the range from about 5 to 200 V. Of course the scale does not conform precisely to the equation $\theta = AV^n$, and it is impossible to use one scale for two ranges in a voltmeter which has to meet an overall accuracy guarantee. However, the correspondence is quite close enough for many practical purposes,

* *Journal I.E.E.*, 1928, vol. 66, p. 893, Table 6.
† See S. E. A. Landale: *Proceedings of the Cambridge Philosophical Society*, 1929, vol. 125, section 4.

* Bibliography, (8), p. 664.

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and the possibility of obtaining an enormous range, at the sacrifice of some accuracy, is worth remembering. We are confronted by the same difficulty when using the systems of Figs. 3 and 4 as diodes, for though the calibration is linear, it is a line which does not pass through the origin. By experiment with either instrument we have $\theta = a(-5 + V)$ from 10 volts upwards.* Suppose that θ is full scale for 100 volts and the galvanometer is shunted so that it is again full scale for 200 volts. If we now apply 100 volts we shall obtain a deflection corresponding to 51·2 *V*, so the two scales disagree at that point by 2·4 per cent. Alternatively, if we make 100 volts applied to the upper range produce the deflection marked 50 on the dial, then 200 volts will produce the deflection marked 97·5 V; the instrument will then be 2·5 per cent low at full scale. If the two ranges are made to agree at about $\frac{3}{4}$ scale and if the high range is never used unnecessarily, then we could use only one scale and never have an error of more than $\pm$ 1·2 per cent. Such a makeshift would be intolerable in a voltmeter which had to meet a general guarantee of accuracy, but it would be no serious disadvantage to an intelligent operator.

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* Bibliography, (11), Figs. 3 and 9.

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A THERMIONIC VALVE POTENTIOMETER FOR AUDIO FREQUENCIES.*

By W. S. STUART, M.Eng., Associate Member.

(*Paper received 21st November, 1929.*)

SUMMARY.

A description is given of a method of measuring audio-frequency potential vectors with the aid of thermionic valves. The arrangement used is a modification of one originally given by A. Pagès, and is adapted for tests on circuits such as artificial lines. A method of estimating the admittance existing between the in ut potential terminals is described.

TABLE OF CONTENTS.

Introduction.
Description of the circuit.
Admittance between the potential terminals.
General remarks on the use of the circuit.

INTRODUCTION.

The measurement of audio-frequency potential vectors is conveniently carried out by means of the well-known circuit, described by Larsen,† in which the e.m.f. derived from the secondary coil of a variable mutual inductance is added to the potential-drop across a pure resistance in series with its primary coil. The total potential provided by this arrangement, which is thus of the form $(R + j\omega M)I$, where I is the current carried by the resistance and primary coil, is balanced against the potential to be measured, the detector being usually a telephone. Balance is effected by simultaneous adjustment of the resistance and mutual inductance, the former measuring the horizontal component and the latter the vertical component of the potential vector, with reference to the phase of the current passing through the circuit.

Larsen, in the paper referred to above, has drawn attention to the care necessary to avoid the effects of stray capacities, particularly in connection with the detector. The disturbing effects of body capacity when telephones are used are well known to users of a.c. bridges, and while, in the case of the bridge, capacity effects may be annulled by taking suitable precautions, there is little doubt that in the potentiometer circuit here considered the capacity existing between the detector and earth may seriously affect the accuracy of a reading. The effect is particularly marked when the method is used to measure the potential developed in a circuit of high impedance.

The difficulty is removed if use is made of triode valves, in that, since the detector is placed in the anode circuit of a valve, its capacity is not thrown across any part of the circuit containing the measuring instruments.

A scheme involving the use of two triodes for this purpose is shown in its simplest form in Fig. 1, the triodes having identical characteristics. The potential to be measured is applied between the grid and filament of V_1, while an equal and opposite potential, derived from the combination of resistance and mutual inductance, is applied to the grid of V_2. Balance is effected by adjusting for silence in the telephone which is common to both anodes.

A practical potentiometer based on this principle has been described by A. Pagès‡, in which a mutual inductance with central zero is used, the arrangement thus permitting the measurement of potential vectors lying within $\pm 90°$ of the potentiometer current vector.

In conducting experiments with an artificial telephone line, the author found it desirable to have a potentiometer with which rapid measurements of potential vectors lying in any quadrant could be made, at frequencies up to 4 000 cycles per sec. After experimenting with several circuits of the type referred to above, it was finally decided to adopt one involving three triodes. This circuit, which is described below, is particularly suited to artificial-line measurement but can, of course, be used for the determination of any potential the magnitude of which is not greater than that which the valves can handle.

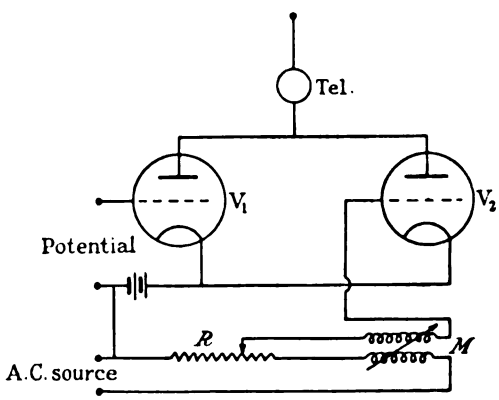

FIG. 1.

DESCRIPTION OF THE CIRCUIT.

It will be seen from the circuit diagram, Fig. 2, that the valve V_2 obtains its grid potential from the resistances R and r, the former consisting of a box with 10 sections and the latter of a slide-wire, while the grid of V_3 is connected to the secondary of the variable mutual inductance M. The primary winding of M is in series

* Reprinted from *Journal I.E.E.*, 1930, vol. 68, p. 769.
† *Elektrotechnische Zeitschrift*, 1910, vol. 41, p. 1039.
‡ *Revue Générale de l'Électricité*, 1926, vol. 19, p. 381

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with R and r, the current passing through this circuit being measured by the thermo-ammeter A. This arrangement enables one end of the secondary winding to be always at earth potential and has been found to minimize the effect of impurity due to capacity between the windings. The advantage is also secured that the switches for reversing the component potentials may be transferred to the anode circuits of their respective valves. Any errors introduced by spurious capacity effects in the grid circuits are thus made definite and may be allowed for if necessary.

The anodes of the valves are connected respectively to three identical coils, L_1, L_2, L_3, mutually coupled by being wound on a ring-shaped iron core. Provided that the mutual conductances of the valves are similar, an e.m.f. applied to the grid of V_1 (i.e. between the terminals E and E') may be balanced by the component e.m.f.'s applied to the grids of V_2 and V_3, when the alternating m.m.f.'s due to L_2 and L_3 exactly balance that due to L_1. Under balance conditions the anode circuits are therefore non-inductive. A fourth winding L_4, connected to a telephone, serves to detect the condition of balance.

It is, of course, essential that the variations of anode current for a given grid voltage, i.e. the mutual conductance, should be the same for each of the three valves. The author attempted to adjust this, in the first place, by varying the filament currents. It was quickly discovered, however, that valves having nominally similar characteristics were actually so different in respect of their mutual conductances that the process of balancing by means of the filament current led, in the case of many valves, to a reduction of the emission below the minimum which could be tolerated. It was finally decided to balance the magnification by means of variable resistances shunting the anode coils as shown. This method was found to be quite satisfactory in practice, since, if the anode coils are tightly coupled, the resistances may be widely different from one another without introducing phase-angle errors.

A variable mutual inductance, shown at M', supplies current from its secondary winding for the circuit under test. In the apparatus as set up by the author the anode coils have 150 turns each, three wires being wound in parallel and uniformly distributed round the ring-shaped core. The latter is built up with 20 laminations of mumetal, having an outside diameter of 5·5 in. and an inside diameter of 3·75 in. The thickness of each lamination is 0·0225 in. The winding L_4 has 500 turns, wound on top of the others. The valves have an a.c. resistance of approximately 3 000 ohms, and with a grid potential of −12 volts pass a current of 12 mA. The anode potential is 120 volts. The resistances R_1, R_2, R_3

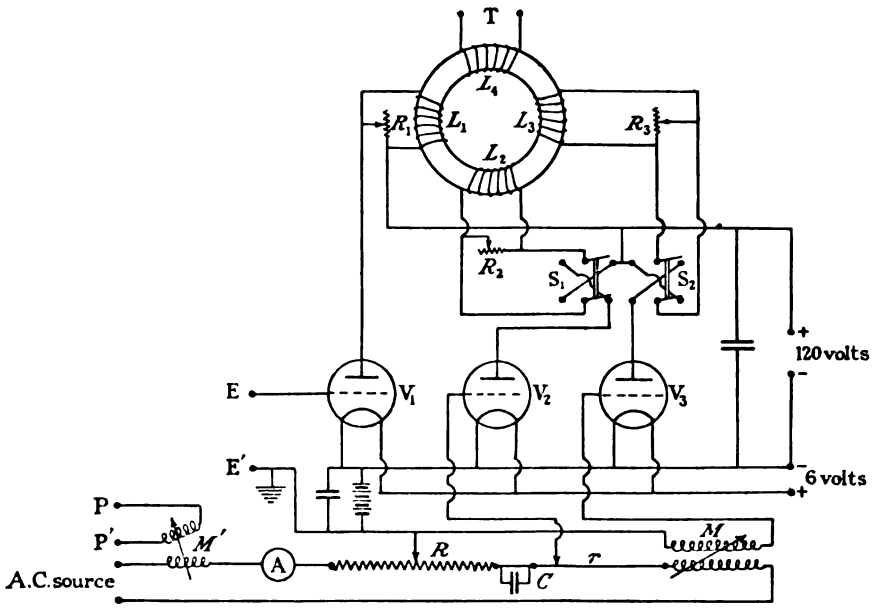

FIG. 2.

are continuously variable and have a maximum value of 5 ohms.

An oscillator of the heterodyne type is used to supply the potentiometer current, which is normally 30 mA. The resistance R consists of 10 sections of 10 ohms each, while the slide-wire r, 1 metre in length, is adjusted also to 10 ohms. The maximum potential which can be balanced by the resistance component is thus 3·3 volts. The mutual inductance M is a standard Campbell inductometer having a maximum value of 11·1 mH.

The employment of a mutual inductance makes the accuracy of the method dependent on an exact knowledge of the frequency. A stretched-wire vibrator,* calibrated by tuning-forks, has been used with success, it having been found possible to determine, to an accuracy of ±0·5 per cent, frequencies which are fundamental or are multiples of the fundamental. An alternative method is to make use of a condenser, shown at C in Fig. 2, inserted in the slide-wire circuit. With R and

* P. K. TURNER: "A Wireless Works Laboratory," *Journal I.E.E.*, 1927, vol. 65, p. 881; also *Proceedings of the Wireless Section*, 1927, vol. 2, p. 194.

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r set at zero, and the filament of V_1 extinguished, a balance can be obtained by adjustment of M. Provided that the phase angle of C is not large, there is no difficulty in arriving at an accurate setting of the inductometer. The frequency is then given by

$$\omega = \frac{1}{\sqrt{(MC)}}$$

Admittance Between the Potential Terminals.

Since one of the most important properties of the 3-electrode valve is the small input admittance of the grid-filament circuit, its utility as a potential-measuring device in association with high-impedance circuits is without question. It is important, however, to know the magnitude of this admittance, so that allowance can be made for it if necessary, e.g. in the measurement of small capacities, and the amount of error introduced by

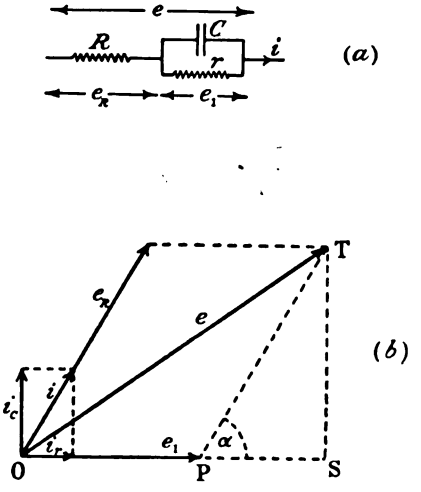

Fig. 3.

neglecting it, in other cases, known. The admittance of the valve V_1 may be measured in the following way:—

The e.m.f. obtained from the terminals P and P′ is measured by connecting these terminals respectively to E and E′. A known pure resistance of the order of 2 megohms is then inserted between P and E and a fresh balance obtained. In this way are obtained the vector voltages which are shown as e and e_1 in Fig. 3(a). In this diagram the valve admittance is represented as a capacity C shunted by a resistance r. The drop across the inserted resistance R is shown as e_R. It is, of course, assumed that the voltage e remains at the same value throughout. Then, referring to the vector diagram, Fig. 3(b),

$$\frac{e_R}{e_1} = R\sqrt{\left[(C\omega)^2 + \frac{1}{r^2}\right]} = \frac{R}{r}\sqrt{[(C\omega r)^2 + 1]}$$

or

$$\frac{(C\omega r)^2 + 1}{r^2} = \frac{1}{R^2}\left(\frac{e_R}{e_1}\right)^2$$

and, since $\tan\alpha = C\omega r$,

$$\frac{\tan^2\alpha + 1}{r^2} = \left(\frac{e_R}{Re_1}\right)^2 = \frac{1}{r^2\cos^2\alpha}$$

or

$$r = \frac{Re_1}{e_R\cos\alpha}$$ and in the figure is $R \cdot \frac{OP}{PS}$

also

$$C = \frac{\tan\alpha}{\omega r} = \frac{\tan\alpha \,.\, e_R\cos\alpha}{\omega Re_1} = \frac{e_R \,.\, \sin\alpha}{\omega Re_1}$$

and in the figure is $\frac{1}{\omega R} \cdot \frac{TS}{OP}$

The values of C and r may thus be conveniently obtained by plotting on squared paper the vectors e and e_1 as given by the potentiometer readings. In performing this test on various valves in the position V_1, it was found that the holder and also the cap of the valve contributed in no small degree to the total admittance.* The following figures show the results of tests, at 2 000 cycles, on a valve having the characteristics referred to above. The resistance R used in the grid circuit was of the grid-leak type and had a value of 1·84 megohms.

	C	r	cos α
	μμF	MΩ	
Valve with cap and holder	22·7	31	0·11
Valve with cap only ..	14·6	54	0·10
Valve without cap and holder	12·2	92	0·07

These figures show the desirability of dispensing with the ordinary mounting of the valve, and the apparatus in its final form now has the valve V_1 supported by means of the bulb, so that direct connection can be made to the grid lead which comes through the glass. The valves V_2 and V_3 are held in ordinary holders.

General Remarks on the Use of the Circuit.

In the setting up of the apparatus it is important to guard against the influence of stray magnetic fields. Unless care was taken to keep these components well separated, it was found that a considerable amount of coupling took place between the inductances M and M' and the anode coils, notwithstanding the toroidal shape of the latter.

The anode shunting resistances are set in the following way:—The grid of V_1 is joined to that of V_2, and a potential is applied to both from the resistance box R. With the switch S_2 open, and S_1 closed in the "first quadrant" position, R_1 and R_2 are adjusted until silence is obtained in the telephone. The grid of V_1 is then connected to that of V_3 and a potential applied to both from M, the switch S_1 now being open and S_2 closed in its "first quadrant" position. Adjustment is now made to R_3 for silence. The adjustment is quite critical, and these rheostats should therefore be capable of continuous variation. Care should be taken to earth the common negative connection to the batteries as shown in Fig. 2.

When taking a reading of potential, the quadrant in which the vector lies may be determined from the position of the switches. It will be evident that when these switches are both in the "first quadrant" position the connection to the anode coils should be such that the

* L. Hartshorn: *Proceedings of the Physical Society of London*, 1927, vol. 39, p. 108; also G. W. Sutton, *ibid.*, 1927, vol. 40, p. 14.

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magnetic effects of L_2 and L_3 are additive, while both are in opposition to L_1. As in the case of the valve voltmeter, the circuit connected to the input terminals must be a conducting one, in order that the grid-bias voltage may be applied to V_1. It will be found desirable to use a bias battery of rather less voltage than is customary when the valves are connected in an ordinary amplifying circuit, since in the present case the external anode impedance is almost zero.

The by-pass condensers shown in the circuit diagram are preferably of large size in order to guard against the effects of unduly high resistance in the batteries.

It has been found possible to read values of potential, over the greater part of the range of telephonic frequencies, with a possible error of 1 mV on either scale. At frequencies below 200, on account of lack of sensitivity in the telephone it will be found advisable to employ a slide-wire current of greater value than the normal, or alternatively to use an amplifier in the telephone circuit.

The author wishes to express his indebtedness to the authorities of the University of Leeds for placing at his disposal workshop and testing facilities, and for permission to describe the apparatus.

MEETINGS OF THE WIRELESS SECTION.

82ND MEETING OF THE WIRELESS SECTION, 5TH MARCH, 1930.

Captain C. E. Kennedy-Purvis, R.N., Chairman of the Section, took the chair at 6 p.m.

The minutes of the meeting of the Wireless Section held on the 5th February, 1930, were taken as read and were confirmed and signed.

A paper by Mr. H. J. Lucas, Associate Member, entitled "Some Developments of the Piezo-Electric Crystal as a Frequency Standard" (see page 151), and a paper by Mr. J. E. P. Vigoureux, M.Sc., Graduate, entitled "The Valve-Maintained Quartz Oscillator" (see page 41), were read and discussed.

The meeting terminated at 7.30 p.m. with a vote of thanks to the authors, which was moved by the Chairman and carried with acclamation.

83RD MEETING OF THE WIRELESS SECTION, 2ND APRIL, 1930.

Captain C. E. Kennedy-Purvis, R.N., Chairman of the Section, took the chair at 6 p.m.

The minutes of the meeting of the Wireless Section held on the 5th March, 1930, were taken as read and were confirmed and signed.

A paper by Mr. R. H. Barfield, M.Sc.(Eng.), Associate Member, entitled "Recent Developments in Direction-Finding Apparatus" (see page 169), was read and discussed.

The meeting terminated at 7.40 p.m. with a vote of thanks to the author, which was moved by the Chairman and carried with acclamation.

84TH MEETING OF THE WIRELESS SECTION, 7TH MAY, 1930.

Captain C. E. Kennedy-Purvis, R.N., Chairman of the Section, took the chair at 6 p.m.

The minutes of the meeting of the Wireless Section held on the 2nd April, 1930, were taken as read and were confirmed and signed.

The Chairman announced that the following nominations had been made by the Wireless Section Committee to fill the vacancies which would occur on the Committee on the 30th September, 1930:—

Chairman: C. E. Rickard, O.B.E.

Members of Committee: G. W. N. Cobbold, B.A., B. E. G. Mittell, G. Shearing, B.Sc., and E. H. Shaughnessy, O.B.E.

In the event of a ballot being required for the new Committee, Messrs. E. W. Moss and F. Tandy were appointed scrutineers.

A paper by Messrs. P. P. Eckersley, Member, and N. Ashbridge, B.Sc., entitled "A Wireless Broadcasting Transmitting Station for Dual Programme Service" (see page 193), was read and discussed.

The meeting terminated at 7.20 p.m. with a vote of thanks to the authors, which was moved by the Chairman and carried with acclamation.

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1930.

EXPLANATION OF ABBREVIATIONS.

(P) indicates a reference to the general title or subject of a paper or address.
(p) indicates a reference to a subject dealt with in a paper or address of which the title is not quoted.
(D) indicates a reference to a discussion upon a paper or address of which the general title or subject is quoted.
(d) indicates a reference to a subject dealt with in a discussion on a paper or address of which the general title is not quoted.

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The Institution of Electrical Engineers
VOL. 6.

WIRELESS SECTION: CHAIRMAN'S ADDRESS*

By C. E. RICKARD, O.B.E., Member.

(Address delivered 5th November, 1930.)

I feel that my first duty this evening is to express my great appreciation of the honour the Wireless Section has done me in electing me to serve as its Chairman for the present Session.

When I look back on the line of distinguished men who have preceded me and on the splendid work that has been accomplished under their guidance I feel that the high standard set by them will be very difficult for me to emulate.

A few years ago one of our past Chairmen remarked that the Wireless Section had not been in existence long enough to have a tradition, but since then we have built up a tradition in which we may take pride. We are now entering our second decade of existence full of vigour, with a membership of 529, backed by a splendid record of interesting papers and discussions in the past, and well-attended meetings.

During the past Session the Council has approved the extension of the activities of the Section to include under-water sound signalling and depth-sounding, communication by infra-red or light rays, and electrical recording and reproduction of sound, as subjects appertaining to the Wireless Section. This additional scope should add further interest to our activities, and we hope that very interesting discussions and exchange of information will develop in due course.

Coming now to the subject of my Address this evening, I think I may safely assume that when inviting me to take this chair you had in mind my association with the development of commercial wireless telegraph communication and the manufacture of wireless transmitting and receiving gear for commercial services, and it is accordingly that part of the wireless art with which I propose to deal.

Speaking of wireless as it strikes the popular imagination, I noticed a few weeks ago that one of our leading daily papers, referring to the recent very interesting and successful public exhibition at Olympia, spoke of wireless as though it had been in existence for only 10 years, which made me wonder what we had been doing during some 20 odd, " pre-wireless " years.

There is no doubt that the advent of broadcasting, to which the newspaper referred, gave a great impetus to the study of radio engineering, and the development of this science in all its branches is proceeding by leaps and bounds. Not all of the inventions or improvements relating to broadcasting are applicable to wireless telegraphy, but there is no doubt that the great advances made in broadcasting as a pub'ic service have placed at the disposition of the wireless telegraph and telephone engineer a vast selec'ion of valves and circuits, and a variety of technique which would not otherwise have been developed.

MARINE COMMUNICATION.

Marine communication was the first form of wireless telegraphy to be developed on a commercial basis and, in spite of its powerful rivals in the public imagination, I am still of the opinion that it is the most important application of wireless, for the simple physical reason that wireless supplies a means of communication with moving objects, particularly with ships and aircraft, which can be achieved in no other manner. The contribution of wireless to the safety of life and property, and its assistance to navigation, are so obvious and so well demonstrated that should there arrive, at some future time, any danger to mobile communication, due to ether congestion, in so far as the essential safety of life at sea or in the air is concerned, I feel certain that international opinion would continue to give marine communication the first consideration, and other forms of wireless communication would be called upon to give way.

These ideas are of course ever present in the minds of the international legislators who frame the rules and regulations of radio-electric communication, and who also demand that marine wireless telegraphy shall progressively improve and keep in step, so far as is economically possible, with modern progress.

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Commenting upon the progress of marine communication we may say that about half (approximately 14 000) of the world's ships are fitted with wireless apparatus. Of these about 10 per cent are fitted with valve transmitters and the remainder with spark only. Spark transmitters of 100 watts input power are considered to provide the best form of emergency apparatus, probably on the principle that the most strident and hoarse cry for help is likely to attract the most attention. Since the 1st January, 1930, in accordance with the Washington Regulations, no new spark transmitters of greater power than 300 watts input power are permitted to be installed, and from now onwards the percentage of high-power spark stations on board ship must of necessity decrease steadily.

The use of short waves for very long-distance communication between ship and shore is being recognized, and the number of ships fitted with short-wave apparatus is steadily increasing, the present-day estimate being about 150 ships fitted with transmitters and three times that number with receivers only.

The fitting of automatic alarm devices on board certain ships where a permanent watch is not kept has been carried out in accordance with the Safety of Life at Sea Convention and there are now approximately 1 200 ships fitted with this apparatus which enables a ship's operator to be called to his post, when he is not on watch, by the ringing of a bell actuated by the reception of the alarm signal which nowadays precedes the S O S call.

There were also some 450 to 500 ships' lifeboats fitted with wireless up to the end of last year.

Apart from the large number of vessels which are compulsorily fitted, in accordance with the regulations of the Safety of Life at Sea Convention, or of the laws of the countries with which, or under whose flag, they trade, there are an increasing number of vessels which carry wireless apparatus although under no legal obligation to do so. In particular I refer to vessels which do not carry passengers and are of less than 1 600 tons gross tonnage. It is obvious in these cases that wireless is carried solely for its utility in the navigation and exploitation of the ship. Among these are a large number of trawlers, whalers and other vessels engaged in the fishing industry, who find that wireless apparatus for communication or for direction-finding purposes is a valuable aid to their industry, and the equipment of vessels of this kind is rapidly on the increase.

So far as marine direction-finding is concerned, the wireless direction-finder is now recognized as a navigational instrument of such importance that at the International Conference for the Safety of Life at Sea, held in London last year, an agreement that all passenger vessels of 5 000 tons or over must be fitted with direction-finders was signed by 18 nations and is to come into force as from the 1st July, 1931.

According to the best estimates available there are at present well over 3 000 vessels equipped with direction-finders, or over one-fifth of the total number of vessels carrying wireless.

Side by side with the ship direction-finder, the installation of automatic wireless beacons around the coasts of the leading maritime countries of the world has made rapid progress. The function of the wireless beacon is to send out a recognized signal at convenient intervals, purely for the purpose of enabling ships fitted with direction-finders to take their bearings and thereby to find their exact position when approaching the coast. Such stations work on wave-lengths between 950 and 1 050 metres and are distinguished by their call signals, the form of signal emitted, and by their musical notes.

At present there are in operation around the coasts of Great Britain and Ireland 15 automatic wireless beacons. Five more are in course of construction, and the number spread over the world is estimated to be about 200.

Rotating Beacons.

As a further aid to navigation an important development has been that of the rotating beacon, to which the Radio Research Board has devoted a good deal of attention.

The results of systematic tests showed that ships and aircraft fitted with ordinary receiving apparatus could obtain their bearings from rotating beacons with sufficient accuracy for navigational purposes, subject to the usual night effects. For small coasting and fishing vessels and for aircraft purposes the rotating beacon has important advantages, in the former case because it avoids for the small shipowner the cost of the ship direction-finding apparatus, and in the latter because the aircraft direction-finder solution is not entirely complete or satisfactory, and there are still some difficulties and limitations to be overcome in certain cases; moreover, every ounce of extra weight carried is an important consideration.

Adcock Direction-Finder.

Shore direction-finding stations as an alternative to rotating beacons or direction indicators have a considerable importance in the field of mobile communications, and there are a number of stations in service both for marine and for aircraft services.

Most of such stations have up to the present been equipped with apparatus of the Bellini-Tosi type, but recently a very notable advance has been made by the development of the Adcock principle, which enables night errors to be eliminated. In this field important investigations have been made by the Radio Research Board.

Parallel work on somewhat different lines had been carried out by the Research Department of the Marconi Co. on the directional characteristics of incoming radiation, on *short* wavelengths. Following their achievements in this field, they decided to carry the technique of their short-wave Adcock direction-finder into the field of medium wavelengths on which radio navigation is carried out. Certain modifications to the method of screening the horizontal leads from the vertical aerials became necessary, as for an accurate navigational instrument the condition of non-radiating screens had to be allied with close electrical symmetry. This was successfully accomplished, and prolonged tests with the new direction-finder showed the possibility, on wavelengths of the order of 1 000 metres, of obtaining a

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night accuracy of 2° to 3°. This result is of the highest importance, when it is realized that hitherto it has not been possible, with practical commercial instruments, to give an absolute guarantee of their accuracy on continuous waves during the night period. This has an important bearing on aircraft navigation.

Comparative tests recently made with aircraft after sunset, during night-effect conditions, over a distance of more than 50 miles, showed that the limit of error was ±3° with the Adcock arrangement, whereas with the Bellini-Tosi method only one-third of the observations made gave bearings of an accuracy within ± 3°, and errors were shown up to approximately 90°. In other words two-thirds of the Bellini-Tosi bearings had errors of ±5° or more.

The very convincing results obtained by the development of the Adcock principle into an instrument of a practical form, which on medium wavelengths can be guaranteed to be free from night-effect errors, is considered to be so important and so definite an advance that it is thought unlikely that future coast or aerodrome stations will be fitted with Bellini-Tosi aerials.

On the other hand, the rigging and top hamper of seagoing vessels is a definite obstacle to the application of the Adcock principle on board ship, and it is considered very unlikely that the Adcock aerial will supersede the rotating-loop or Bellini-Tosi aerials for marine direction-finding purposes.

Equi-Signal Radio Beacons.

Another direction-finding problem peculiar to aircraft is that of "homing," or flying to and approaching an aerodrome. In this case a pilot does not so much desire to know his position as defined by bearings on two fixed stations, which may perhaps involve reference to a map, as to know in particular whether he is flying direct to his destination. Since 1920 the American Bureau of Standards has been developing a method of directional transmission which would serve this particular purpose.

The underlying principle of the American radio beacon system may be briefly stated as follows: Two directive aerials are placed at an angle with each other. Along the line bisecting this angle the radiation from each aerial is equal. At any other point radiation is stronger from one than from the other. If, therefore, the separate radiations of the two directional aerials can be clearly designated, it becomes possible by observing the combined radiation of the system to follow the course along the bisector referred to.

A very practical form of such a beacon was produced, known as an aural equi-signal radio beacon. It consisted of a Bellini-Tosi aerial system excited through two separate radiogoniometers set at 90° or less to each other. The radiation via the two radiogoniometers was distinguished by the letters "N" and "A", and the keying was interlocked in such a manner that equal reception of the two radiations combined to form a continuous dash. This dash indicated to the receiver that he was placed exactly on the line bisecting the angle between the two figure-of-eight diagrams of radiation. Displacement to the left or right of this line was indicated by reception of either the letter "N" or "A." Several such beacons are now in operation in the United States for point-to-point navigation of aircraft.

The Bureau of Standards of America have also made experiments with beacons in which the separate directional radiations are designated by different modulated frequencies. The output circuit of the receiver developed to work in conjunction with such beacons is fed to two reed indicators, the first of which is resonant to one frequency of modulation, and the second to the other. If, therefore, the receiver is placed on the line bisecting two such radiations, both reeds vibrate equally, whilst a displacement to the right or left of the line coincides with a reduction in amplitude of vibration of one reed and an augmentation of the other, and vice versa.

Ship Telephony.

Ship-to-ship telephony is in common use between trawlers and whalers and other types of vessels, and ship-to-shore telephony has been successfully carried out on many occasions previously, but perhaps the most interesting development in marine communication during the past twelve months has been the combined effort made by the British Post Office on this side of the Atlantic, the American Telephone and Telegraph Co. in the United States, and the Marine Communication Companies and their associated manufacturing Companies to try out a public ship-to-shore radiophone service. So excellent and comprehensive an address was given a few months ago by Sir Thomas Purves to the London Students' Section that there is little I can add on this subject.

Compared with transatlantic telephony, the obvious transmission handicaps of telephony from shore to ship-board in the Atlantic are (1) the economic limit to the power of the ship's transmitter, (2) the lack of the multiplying factor of beam aerials both for the transmitter and the receiver on board, and (3) the continually changing distances over which it is required to work. The small space available for the installation not only limits the power of the transmitter but has a bearing in some cases on the number of wavelengths it is practicable to employ to meet the changing conditions of distance, of night and day working, and of the seasons of the year.

Another drawback is the fact that, owing to the changing direction of the ships from the shore stations, the fullest advantage of directional beam working cannot be taken at the shore end. There are further difficulties due to the fact that those ships with which a ship-to-shore telephony service is most desirable are those with which it is also necessary to maintain the best telegraph service, which entails simultaneous working on short-wave telegraph sets for arbitrage or Stock Exchange services, medium long-wave working on 1 875 to 3 000 metres for long distances; 600 to 800 metres working on interrupted continuous waves for normal ship communication purposes, and reception of news on long waves from Rugby and other news-disseminating stations. All of these things are desirable and are accomplished to a certain extent, but the problem of interference to the short-wave telephony service from the long-wave ship transmitter on continuous or interrupted continuous waves is serious.

The type of continuous-wave transmitter that it is

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economically possible, or even practicable, to install on shipboard is a generator of numerous harmonics, and it requires co-operation between telephonist and telegraphist on board to avoid interference, and a certain amount of juggling with wavelengths to prevent a harmonic of the long-wave set interfering with telephony. Perhaps the most difficult problem which has arisen is the fact that the powerful radiation from the short-wave transmitting aerial generates oscillations in all loose conductors about the ship. These conductors re-radiate and, given the vast number of possible modes of vibration, there is practically certain to be interference with the incoming telephony wave. Bonding of stays and conductors has been resorted to, but this is a difficult job and it is usually impossible to trace the particular conductor or combination of conductors causing the interference. Some ships are worse than others in this respect. These parasitic re-radiations behave somewhat similarly to atmospherics, and the best solution appears to be to suppress the cause, that is to say to suppress, or rather to reduce greatly, the carrier of the transmitted wave during reception.

Reviewing the situation, I think we must look upon the work done up to the present with regard to ship-to-shore telephony as a series of successful experiments which have shown the direction in which we must proceed. I believe it can be said to be established that 4-wire working at the ship end is merely a stepping stone and that we must face the expense of 2-wire working, and the use of a telephone terminal equipment on board, if a satisfactory service is to be attained. With regard to the number of wavelengths required—a problem which affects the manufacturer and legislator concerned with the repartition of the ether—it is at present considered that 5 wavelengths are advisable if a continuous service is to be obtained, day and night, winter and summer, and at the varying distances from land over which communication has to be made. Such wavelengths as are being catered for at the moment by the Rugby and American shore stations are 17, 23, 34, 63 and 90 metres approximately. Experience is not sufficient to establish these definitely, and it may be that 26½ metres as a compromise between 23 and 34 and 75 metres as a compromise between 63 and 90, will be found sufficient for commercial purposes when further experience is gained, or some other compromise may be found sufficient for ordinary purposes, thus reducing the wavelength requirements to three.

Slightly separated wavelengths are required for the shore stations in America and Europe to allow simultaneous conversations to be carried on from either side of the Atlantic, and ship stations require to be able to make slight adjustments to each of their waves to enable them to call either Rugby or New York.

The selection of wavelengths for the ship telephony service is complicated by the presence on board ship of short-wave installations working telegraphy; and in fact the selection of wavelengths and the organization of all short-wave services is becoming so delicate a matter that it is to-day a specialist's job.

It seems certain that the study of ship-to-shore telephony organization, and frequency allocations, will provide an interesting and, at the same time, arduous task for the experts of the International Technical Consulting Committee, who meet again next year at Copenhagen.

Design and Development of Portable and Semi-Portable Apparatus for Mobile Services.

Transmitters.

In the field of wireless engineering concerned with small- and medium-power general-purpose transmitters, activity has centred chiefly in the last few years in the development of circuits to make possible the application of short waves. Naval, military and aviation authorities, as well as civil authorities responsible for the maintenance of communication between small isolated centres, have been quick to appreciate the great saving in power and cost of transmitters made possible by the use of short waves, and, further, the possibility of communicating over distances quite outside the range of medium-wave transmitters of a practicable power, size and economic cost.

In the class of transmitter under consideration the types vary from pack sets employing hand-generators to fixed or mobile stations of the order of 3 kW aerial power, but in every case the prime requirements are a large choice of transmitting wavelengths and simplicity of operation, the latter being of the utmost importance since this class of transmitter has frequently to be operated by entirely unskilled personnel.

Arising out of these requirements the problem most difficult of simple solution, from about 75 metres downwards, is the tuning of the aerial circuit.

Two methods are in general use:—

(*a*) To employ a variable length of aerial, using an insulated winch, and so to adjust its length that the wavelength of aerial plus loading and coupling coils is one-quarter or three-quarters of the transmitted wavelength.

(*b*) To use a fixed length of aerial and to provide at its base a variable impedance capable of terminating correctly the aerial for any distribution of standing waves along its length.

Method (*a*) gives better average radiation but is impracticable in many cases, e.g. a submarine; while method (*b*) lends itself to extreme simplicity of adjustment, and for this reason, and also because it is simpler to rig, it is more frequently employed, in spite of poorer average radiation.

One important aspect of the introduction of short waves into this class of transmitter is that whereas previously, on medium wavelengths reliable telephonic communication has been possible with simple self-oscillators, this is not the case on short wavelengths, due to distortion during propagation resulting from variations in the carrier frequency caused by modulation.

Most short-wave transmitters of this class, therefore, employ only C.W.* and I.C.W.† telegraphy, but for some services a demand exists for telephonic communication, and thus the difficulties of design have been greatly increased by having to provide a master oscillator the

* Continuous-wave. † Interrupted continuous-wave.

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frequency of which will be unaffected by modulation of a subsequent amplifying stage. Also, international regulations as to constancy of frequency now make it necessary to employ master oscillators having a high degree of constancy for certain services catered for by this class of transmitters.

A modern example of a transmitter designed to cover all these varied requirements is a crystal-controlled set the main data of which are as follows:—

Wave-range ..	16 to 75 metres.
Master oscillator ..	Choice, by external switch, of a valve oscillator giving continuous wave-range, or three short waves by crystals. Valve oscillator and crystals enclosed in thermostatically controlled box. By suitable doubling and trebling in subsequent stages, valve oscillator covers wave-range in one sweep of variable condenser.
Aerial power ..	C.W. 350 watts; I.C.W. and telephony 120 watts.

Instantaneous selection of C.W., I.C.W. or telephony by 3-position switch.

Keying	By self-absorber method, up to 150 words per min.
75 % modulation ..	On telephony—sensibly linear.
Aerials	A single length of aerial can be employed for whole wave-range, but naturally, when working wavelengths have been decided upon, aerials would be used having good radiation at those wavelengths.
Carrier frequency	Constant to 1 in 10 000.
Time to change from one frequency to another	5 minutes in worst cases.
Size	5 ft. × 5 ft. × 18 in. deep.

At the other end of the scale is a portable set for C.W. and I.C.W. telegraphy only, employing a self-oscillator, wave-range 20 to 60 metres, using a hand-generator as a source of power, and capable of operation by semi-skilled personnel.

Modern aerodrome ground station transmitters present special problems, the chief of which is the selection by remote control of any one of six spot wavelengths in the waveband 400 to 1 600 metres.

Still another tendency in all transmitters of this class is the elimination of batteries for auxiliary supplies, as for keying, grid bias, etc. Small rectifiers—of the metal type where possible—are used to provide these supplies from the filament alternator.

Screen-grid transmitting valves have opened up new possibilities in the simplification of circuits, since they dispense with the necessity for balancing interelectrode capacitances in multi-stage transmitters. Against this advantage must be set their lower efficiency and the requirement of a screen-grid voltage supply. In small- and medium-power transmitters, up to, say, ½ kW aerial power, the latter is important, since it involves either a separate h.t. supply or a potentiometer across the main h.t. supply taking an amount of power which is uneconomical in proportion to the total power of the transmitter. In many cases, however, the resultant circuit simplification more than justifies the use of such valves in general-purpose transmitters.

Receivers.

Coming now to general-purpose receivers, the two most important influences in recent years in their progressive development have been the introduction of short waves and the arrival of the screen-grid valve. For receivers in this class the prime requirements are a wide wave-range and simplicity of operation.

Development has been along two main lines:—

(*a*) Receivers with low damping input circuits, followed by high-gain multi-stage screen-grid valve amplifiers, with reaction definitely excluded. Separate oscillators for C.W. reception, tuned or untuned low-frequency magnification at will, or

(*b*) Receivers with one stage of screen-grid valve amplification, having the aerial directly coupled to the grid circuit, followed by a detector with reaction capable of self-oscillation for C.W. reception. Low-frequency amplification as for (*a*), i.e. tuned or untuned.

Type (*a*) is confined to medium and long waves, and development has resulted in the use of at least three stages of high-frequency amplification with small gain per stage, for the reasons that such an arrangement does not depend on special care in design or testing to ensure stability, and also that the resultant high damping of this circuit enables the condensers to be ganged.

Type (*b*) fulfils the requirements of a great many services and is used on short, medium and long waves. Below about 150 m such receivers usually employ plug-in coils, to avoid stray capacitances as far as possible, but above this wavelength all coils are contained in the receiver and the range required is selected by a switch.

A modern example of a receiver in category (*b*) has a wave-range of 150 to 3 000 m. Circuit development has made possible the complete ganging of the two tuning circuits, in spite of the fact that an aerial is coupled to one and reaction applied to the other and that both are low-loss circuits, so that the tuning controls comprise a 4-position range switch, one tuning handle —calibrated directly in wavelengths—and a reaction control. A drastic volume control is required on this class of receiver, because mobile service conditions frequently imply working at short ranges from powerful transmitters. Such a volume control should increase high-frequency selectivity whilst reducing high-frequency gain.

Aircraft Picture Transmission.

The problem of transmitting military intelligence by means of picture telegraph apparatus between moving aircraft and a distant ground station has been investigated and the results obtained show that the methods avail-

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able are quite practical. One such apparatus recently developed possesses the following features:—

(*a*) Independent synchronizing means at the transmitter and receiver.
(*b*) No photo-electric pick-up system at the transmitter.
(*c*) Messages to be transmitted are written in conducting ink or very soft pencil.
(*d*) Messages at the transmitter and receiver remain stationary.

Actual flying experience has shown that it is possible to convey military intelligence between an aircraft and its base over distances up to 150 miles. The normal aircraft wave using a 160-watt transmitter is employed and the speed of transmission is approximately 9 square inches per minute. With the adverse signal-to-noise ratio available on such a circuit it has been found essential to utilize separate synchronizing means at the transmitter and receiver.

the disturbances to the short-wave services will become less and less during the next 5 or 6 years, increasing again until the period of intense magnetic disturbance occurs 10 or 11 years hence, by which time we hope that some method will have been found of eliminating their effects.

As an illustration of the effects of magnetic storms on long-distance short-wave communication I have selected from a series of curves an example for the months of April and May of this year (see Fig. 1). Traffic between England and Australia is sent on 26-m waves, part of the day over the " long route " across the Atlantic and South Pacific Oceans, and part over the " short route " across Asia and the East Indies.

The rounded curve refers to the " long " route, which is the worst and least used, and shows the reduction of hours of high speed and recordable working, as compared with the incidence of magnetic storms.

Fortunately the " short " route is not so badly affected. The Indian and South African wireless circuits are

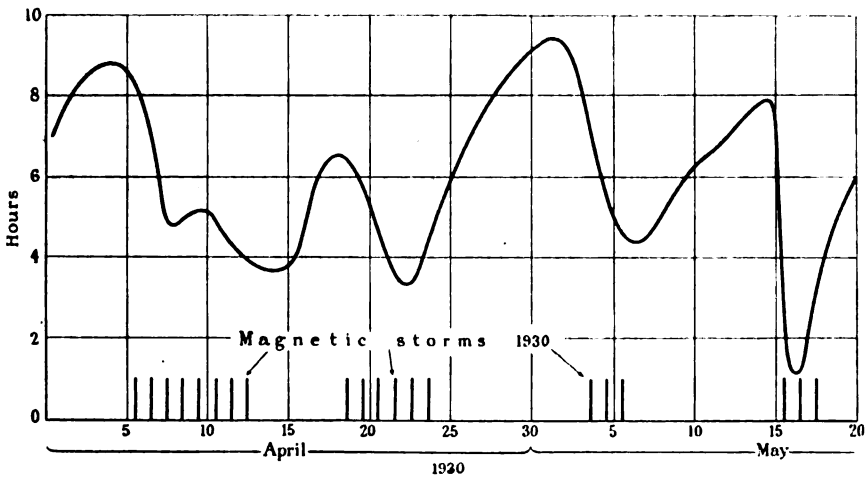

FIG. 1.—Total recordable hours per day on long-route Australian circuit.

Let us turn now from the mobile to the fixed services.

POINT-TO-POINT COMMUNICATION.

The commercial development of point-to-point services has been continuous during the past year. On the telegraph side new circuits are being opened up for communication and the number of competitors against our national services is steadily increasing. Since the War the tendency for each nation to control its own communications becomes continuously stronger, and whereas wireless communication on long waves did *not* constitute a really serious competitor to cable companies, short-wave working goes strongly ahead, and each new commercial wireless station in foreign countries takes traffic and revenue from the older methods of communication.

During the present year, however, the short-wave services have not quite lived up to their reputations of the previous years, and there has been a falling-off in the number of hours of recordability, due to an excessive number of magnetic storms. The year 1930 has been peculiarly severe in this respect, but it is probable that hardly affected at all. It is those circuits of which the great circle between terminals passes near or over the polar regions which suffer most during magnetic storm periods.

These facts point to a solution, i.e. the use of relay stations during fade-out periods in such a way that the great-circle paths between terminal stations and relay stations do not approach the polar regions.

There is some reason to believe that the most suitable wavelength for communication between distant points is considerably shorter during the periods when magnetic storms are prevalent than in the periods of solar quiescence. Investigations are being carried on in many parts of the world in the hope of finding a solution in the use of very short waves. Also it may be stated that during the early days of experimental short-wave working before the beam stations were constructed, wavelengths of 30 and 60 m were found quite suitable for transatlantic day and night working respectively. The beam stations were originally designed for such wavelengths, which were found at a later date to be quite unsuitable.

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There is therefore a possibility that, as time proceeds along the magnetic storm cycle, it may be found necessary to modify the present beam-station wavelengths. There is not much evidence to support this theory, but it is worthy of careful observation.

During the periods of magnetic storms it has been found on repeated occasions that while communication on short waves has failed completely during periods of 12 hours or more, communication on long waves has remained unaffected. Therefore, in spite of the higher initial cost and higher maintenance charges of high-power long-wave stations they still have their uses, particularly for distances up to 3 000 miles, and both the Post Office and Imperial and International Communications Ltd. contemplate additional long-wave stations for their communications with America, the Post Office for telephony and Imperial and International Communications Ltd. for telegraphy.

Point-to-Point Telephony.

The Post Office telephony services from Rugby continue to extend rapidly, and the dream of telephonic intercommunication between the Dominions, Colonies and the Mother Country is in course of becoming an accomplished fact. I do not think I can add anything to the excellent papers and information which have been read and published on radio-telephone services, in particular the lecture given by Mr. G. H. Nash, C.B.E., at the Royal Society of Arts in April last, but perhaps the following remarks will be appropriate from one whose experience was confined to radio-telegraph services until comparatively recently.

Until radio-telephony was developed to the stage of entering into public service, radio communication could hardly be called an exact science. It was, of course, as exact as we could make it, but there is no doubt that the influence of telephone practice on radio practice has been very beneficial. In particular the adoption of the transmission unit (néper or decibel) and of methods of measuring amplification and losses in our apparatus has led to a great advance in technique in all directions; for example, it is now standard practice to specify the gain in the individual parts of a receiver, the shape of the filter curves, and even the gain of beam aerials. Added to this, the habit of measurement and study of transmission phenomena has extended rapidly as a result of the more general use and progress of measuring apparatus and methods.

Development in radio practice both in measurement and in theoretical studies has made it possible to prophesy the behaviour of a proposed radio channel with increasing accuracy. The propagation phenomena in connection with waves from 15 to 40 m have been investigated so that for a given period of the year, time of day, and wavelength employed, it is possible to prophesy with reasonable precision the conditions which will prevail at either end of a radio channel between given points. Such studies have proved of considerable value in the setting up of intercontinental wireless circuits.

Reverting to the influence of telephone practice, its value was exemplified in planning, for example, the Rugby–New York radio-telephone circuit. In operation such a circuit is treated, from the point of view of a telephonist, as if it were a metallic circuit, and in the main the circuit is treated as though it were a repeatered line.

Methods of design employed in telephone practice are now used to an increasing extent on wireless receivers and transmitters, such as rack-mounted equipment, as in repeater stations and alarm indicating devices to show the existence and whereabouts of faults and to provide facilities for testing circuits at any point throughout the equipment, etc. It may be said that the influence of telephone practice on wireless practice, particularly as regards telephony, is due to the conditions under which telephony must work. Since a subscriber has no technical interest or appreciation of any difficulties which may arise, it is essential that there be no discontinuity in his conversation. Under these circumstances faults must be cleared with the minimum possible delay, and all levels throughout the system correctly adjusted and supervised. In pursuit of the telephonists' ideal of uniform signal intensity, a great deal of money has been spent in the effort to perfect a method of automatic volume control which will maintain a uniform level during extremes of fading without causing distortion.

Facsimile Working.

The attainment of facsimile transmission, that is to say the transmission of a message in the form in which it is handed in at the telegraph office, has been the goal of endeavour of administrations and companies exploiting telegraph communication for a number of years, and public services have been open for a considerable time for the transmission of facsimile pictures and messages.

By *wire* telegraphy the problem of picture and facsimile working can be said to be solved to a large extent, but the problem of providing high-speed facsimile services on long-distance short-wave circuits has not yet been solved. Modern facsimile transmitting or receiving apparatus does not impose an upper speed limit on such a circuit, but the intervening space between the transmitter and receiver does impose a very definite limit on the speed at which facsimile messages can be faithfully reproduced. In other words, high modulation frequencies are not possible on such circuits. This limitation is due to the fact that although one pulse or dot may be signalled, the waves so radiated will be reflected from the Heaviside layer a number of times and the distances traversed by the various rays will be unequal. At the receiver end a number of overlapping pulses or dots will be received and the delay between the first and last dot may be more than 0·01 sec.

For example, during tests on 60-m waves between Somerton and Chelmsford, a distance of about 144 miles, the maximum time interval observed between the first and last echo was about 0·005 sec. during the day and 0·0143 sec. during the night. On transatlantic circuits using waves of about 22 m the interval between the first and last impulse varies from about 0·001 sec. to 0·01 sec.

In short-wave radio-telephony the ear and brain do not take notice of the relatively short intervals between these echoes, and as a result telephony, that is to say high modulation frequencies, can be employed without apparent distortion (assuming the terminal apparatus

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to be free from distortion). The facsimile machine is, however, really a photographic oscillograph, and a very clear record of the mutilation is shown by a series of multiple images of a single transmitted impulse. Therefore, in order to achieve successful transmission of facsimile, the speed of working has to be adjusted to the prevailing circuit conditions, i.e. it has to be reduced to such a value that the overlapping impulses merge into a single line and the apparent distortion is thus made negligible.

The following table gives some figures of the speeds of

Linear velocity of exploring spot	Period	Typewriting in 5 letter words per minute
in. per sec.		
3·61	Twilight and night	20
7·22	Daytime	40
10·83	Abnormal daytime conditions	60

working found to be commercially practicable under transatlantic conditions using beam aerials. The wavelength employed was 22 m and the distance 3 000 miles. For those not familiar with the subject it may be mentioned that 10 typewritten words occupy, with standard lettering, a space of approximately 1 sq. in.

Actually the facsimile terminal equipment is capable of reproducing messages at linear exploring spot velocities as high as 120 in. per sec., corresponding to a speed of 66⅔ sq. in. per min., or a typewriting speed of approximately 665 words per min.

My photographic illustration* shows the effect of attempting to transmit facsimile at high speeds. A linear spot velocity of 10 or 11 in. per sec. is the speed I have cited as being the maximum on 22-m waves on the transatlantic circuit, under abnormally favourable conditions. At higher speeds, say 15 in. per sec., distortion begins to show up, at 30 in. per sec. it is very bad and at 60 in. per sec., or the equivalent telegraph speed of 330 words per min., the script is unreadable.

The problem of increasing the speed of facsimile transmission is being attacked by increasing the directivity of the aerial polar diagram in the vertical plane, so as to limit the solid angle of radiation of aerial systems. Thus it is hoped to limit as much as possible the number of rays passing between transmitting and receiving aerials. Furthermore, present theories suggest that on shorter waves, for example 16-m, better results are to be expected than heretofore, and that the maximum echo lag on a 16-m transmission to New York is likely to be only one-fifth of that on 22 m. Results of tests to be carried out on 16 m between England and Montreal were therefore awaited with keen interest. They have just come to hand and, happily, confirm the accuracy of the prognostication.

Reverting to an earlier remark, that the facsimile machine is really a sort of photographic oscillograph, it is interesting to observe that experiments with this apparatus have given unexpected impetus to the study of short-wave propagation phenomena. From the results of tests accurate data have been supplied upon which scientists have been able to build up their theories.

* Not reproduced in the *Journal*.

Developments in Point-to-Point Services and Practice.

The use of directive aerial systems both for transmission and reception on short waves has now become an accepted principle and established practice, and a number of different shapes and forms have emerged. Endeavours are now being made to improve their directive efficiency and to control their polar diagrams both in the horizontal and vertical planes, either by varying the form and disposition of the aerial wires and reflectors, or by inclining them from the vertical or horizontal, as the case may be.

In the interests of economy, the design of a directive aerial becomes a compromise based on the probable earning capacity of the proposed circuit. Economy has been sought in feeder and distribution system costs by using twin open-wire feeders, as opposed to concentric copper tubes. So far as the transmitting station is concerned, radio engineers are perhaps not yet agreed. On the one hand concentric tube feeders are reliable in service and certain in behaviour and efficiency, but their cost is high. On the other hand open-wire feeders are cheap, but their behaviour as non-radiators is more difficult to control and consequently their efficiency is more variable.

At the receiving station the concentric-tube feeder has the valuable advantage of perfect screening, or absence of pick-up from local sources of interference.

With competing services arising in all directions the natural tendency is to reduce initial costs and maintenance charges wherever possible, and recent practice has been to group a large number of transmitters or receivers, as the case may be, on the same site and to locate such sites at a moderate distance from a central office situated at a traffic centre.

A further development is to situate both transmitter and receiver on the same site where possible, of course without sacrificing the advantages of duplex working. This has been done at a number of stations and is contemplated in a number of cases where new stations are to be erected. The principle employed in this achievement is quite simple, i.e. the receiving aerial is placed in the direction of zero radiation from a beam transmitting aerial, or vice versa. In either case a small difference of frequency is quite sufficient to avoid interference between transmitter and receiver.

The type of cyclic fading known to radio receiving engineers as selective fading is being attacked both at the transmitting and receiving ends of the circuit by the so-called diversity methods. Modulating the transmitter output at a tone frequency yields a diverse frequency effect which has been found to improve considerably the general level of signal strength, with the result that since modulation has been introduced there has been a very noticeable increase in the average effective speed of communication on short waves, accompanied by an increase in the average daily number of hours of recordable signals. At the receiving end diversity reception, or simultaneous reception of signals on separate or spaced receiving aerials, is also found to

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be very effective. Large-area directional aerials, both for transmitting and receiving, have a similar effect.

Modulation is not favoured in some quarters, on the ground that it occupies a wider band of frequencies than C.W., but it has the advantage that it improves the reception of distant stations where for economical reasons or lack of space, as on board ship, it is not practicable to employ beam or diversity aerials.

Diversity, i.e. spaced, aerials, are practicable for large receiving stations where a number of circuits and, therefore, receivers are concentrated at one place and can be jointly connected to the same aerials. Generally it may be said that each method has its advantages according to circumstances.

In order to cope with the high signalling speeds required for facsimile keying, for example, and to effect economy in land-line charges between the central radio office and the transmitter, improvements are continually being made in keying methods, and it is becoming the established practice to employ tone sending over the intervening land lines and cables.

As regards installation design, the present tendency is to take 3-phase power from the supply mains and to eliminate all running machinery and batteries by using static transformers and rectifiers to generate all plate, grid and filament supplies.

Much work is also being done on the design of reliable frequency meters for station use, and also in the design of commercial apparatus for the rapid and accurate measurement of incoming waves, based on multivibrator or other fundamental methods.

Any one acquainted with short-wave working or even with the Berne list of frequencies must have observed that the separation between station allocations is becoming smaller and smaller and that the need for frequency constancy is becoming more evident every day. The tendency is therefore not only for the controlling administrations and governments but also for the exploiting companies to set up monitoring stations whereat they can make rapid and accurate measurements of the outgoing frequencies of stations under their control or with which they are in communication, and also of interfering stations. In some quarters a further step has been made and field-strength measuring sets are being used to advise or give warning as time progresses of a change of frequency on any particular channel.

Frequency Control.

The need for constant carrier-frequency emission on all services becomes more and more evident every day. A step in advance from the international point of view was made at the Hague last year, when certain recommendations were made for the permissible tolerances in the various frequency bands, and administrations and manufacturers are in the main endeavouring to conform to them. The standard was not set too high for achievement in the majority of cases, but experience shows that there is a great difference between the measure of constancy a master oscillator is expected to maintain and what is experienced in actual practice. Many a constant-frequency device has given promise in the laboratory of a frequency constancy of, say, 1 in 50 000 or 1 in 100 000, but in service spread over a week or a month its figure of merit has been found to be no better than, say, 1 in 10 000, or less.

It is not sufficient for a master oscillator to keep within a certain percentage of its mean frequency at any given time; it must also not wander from its allocated frequency, that is to say, the official frequency on which it is required to work. A carrier frequency which is stable, but inclined to drift, can be followed at the receiver end until it approaches a neighbouring station frequency and interference arises. On the other hand, an inherently unstable frequency, or in other words a wobbler or scintillator, is fatal to good telephony or broadcasting quality.

The three best-known constant-frequency devices in commercial use are the tuning fork, the quartz crystal, and the valve master-oscillator.

Tuning Fork.—The tuning fork has long been used as a sub-standard of frequency, and many of the frequency-measuring sets employ tuning forks as a basis of measurement which can be checked from time to time.

Where synchronization has been required, as in picture transmission for example, tuning forks have been used for some time. Forks made of elinvar have been found sufficiently stable, without temperature control, to allow of the device being used with considerable success on long-wave transmitters.

For the maintenance of constant frequency with forks of other materials it is indispensable to employ temperature-control apparatus, capable of maintaining the fork at constant temperature to within approximately 0·1 degree F. if the best results are desired. Under these circumstances the results obtained have been remarkably good. The 10 tuning-fork devices employed by the British Broadcasting Corporation to control the transmission of the relay stations on a common frequency of 1 040 000 are, I understand, still maintaining their high reputation for constancy after many months in service, and I believe their constancy can still be defined as about $\pm$ 1½ in 100 000.

The principle of tuning-fork control is being extended to high frequencies, i.e. short waves, where the difficulty of avoiding undesired modulation of the final frequency is more acute. This trouble has led to the use of a chain of doubling stages in preference to any other method.

Developments in the method of maintaining a fork in vibration are in the direction of minimizing the effects of supply voltage fluctuation on the frequency, and the elimination of any possibility of unwanted degrees of freedom.

Quartz Oscillator.—The study of the quartz oscillator has revealed the fact that if the section is suitably dimensioned the temperature coefficient of the plate and its holder can be made positive or negative as desired. By adjusting the dimensions it is possible to change the temperature coefficient from one sign to the other at the expense of some slight change in the frequency due to the cutting. Further, by very careful adjustment of the dimensions it has been found possible to produce quartz oscillators of a given frequency the temperature coefficient of which is of the order of $\pm$ 2 or 3 parts in 1 million per degree C. change of temperature.

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Unfortunately, whilst this result has been obtained on crystals of a frequency as high as 250 000, at still higher frequencies the crystal itself is thin, and its behaviour becomes spasmodic.

With crystals of ideal dimensions laboratory tests have shown that changes of valve voltages of the order

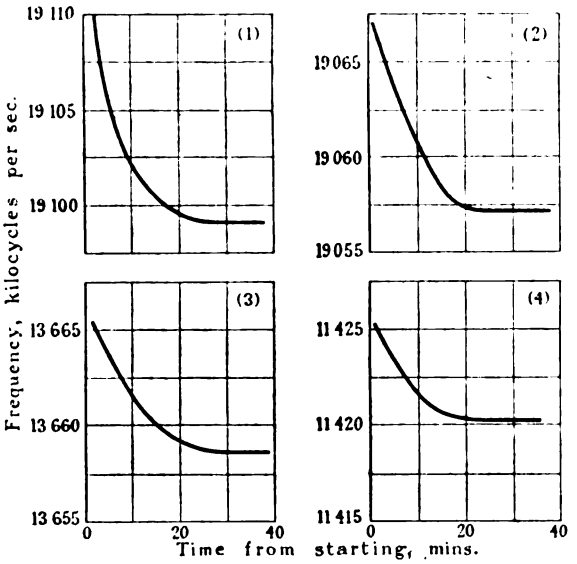

FIG. 2.—Initial frequency drift in beam transmitters.

Curve (1).—GLW. $\lambda = 15{\cdot}707$ m (19 101 kilocycles).
Curve (2).—GLG. $\lambda = 15{\cdot}740$ m (19 060 kilocycles).
Curve (3).—GLL. $\lambda = 21{\cdot}962$ m (13 660 kilocycles).
Curve (4).—GLY. $\lambda = 26{\cdot}269$ m (11 420 kilocycles).

of $\pm$ 5 per cent have produced such small frequency-changes that to make an absolute measurement is a matter of great difficulty.

Some measurements of three quartz-driven stations working on a common wave of about 1 318 kilocycles per sec. have shown that between A and B stations

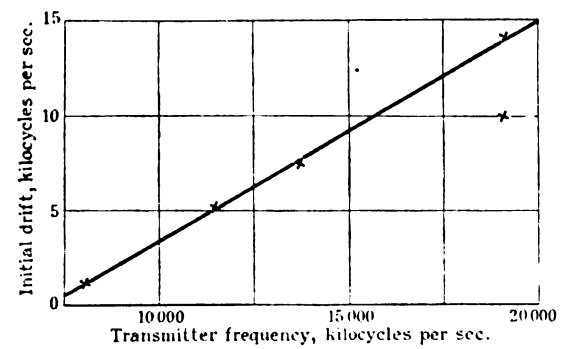

FIG. 3.—Initial frequency drift in beam transmitters.

there was a frequency change of from 0 to 16 cycles per sec., whereas between A and C stations there was a frequency change up to 80 cycles per sec. These changes are therefore of the order of 1 to 5 in 100 000.

Valve Master-Oscillators.—With this type of constant-frequency device the main difficulties to be overcome are the variations of frequency due to dimensional changes of the oscillatory circuit, that is to say, condenser and inductance, due mainly to changes of temperature or flow of the material, and changes of supply voltages to the valve filament, anode and grid. Several circuits are known which take care of the valve problem, but the attainment of stability of the oscillation constant of the condenser-inductance circuit over a long period, even when maintained at constant temperature, is a problem of considerable difficulty.

Mr. C. S. Franklin has made a special study of short-wave constant-frequency devices and has developed one in which the condenser-inductance circuit is of particularly robust construction, of special materials, and in which the principle of automatic correction for temperature variation is applied.

At this point one must distinguish between the transmitter fitted with an ordinary valve drive and the type developed by Mr. Franklin, in which the constant-frequency device is composed of a number of stages, including doublers, on exactly similar lines to the quartz-crystal constant-frequency device, so much so that in

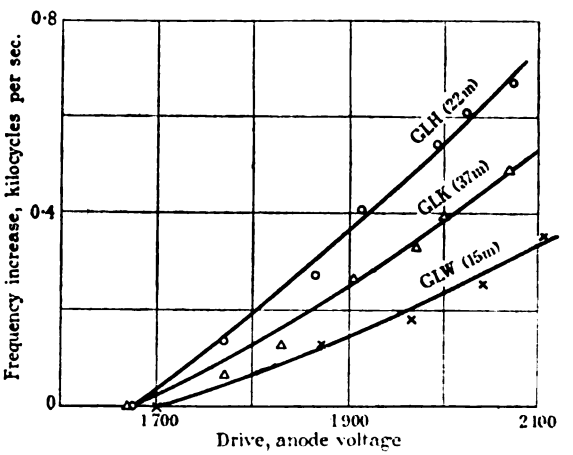

FIG. 4.—Frequency variations with drive anode voltage in beam transmitters.

the new device a crystal may be inserted in the circuit if so desired.

Taking ordinary valve drives first, much can be done in the interests of frequency stability by eliminating vibration, by using accumulators for filament lighting and employing a very constant h.t. supply to the master oscillator, together with the use of as low a power as possible and an isolator stage, and by careful screening of the master-oscillator and isolator stages from the rest of the circuits. To this must be added the use of inductances and condensers of robust and inflexible design, the whole maintained at as even a temperature as possible. Such was the design of the beam station transmitters, the note of which was for a long time recognized as the most constant of those in regular service.

Valve drives of the normal type suffer, however, from drift due to starting-up conditions (principally temperature variations). The short-wave marine transmitter with or without a master-oscillator is a bad offender in this respect, and when such stations are started up they may drift over 10 to 50 kilocycles per sec. during the first

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half-hour. This constitutes a great difficulty in short-wave ship communication because, as a rule, the normal ship station does not require to work for more than half-an-hour at a time.

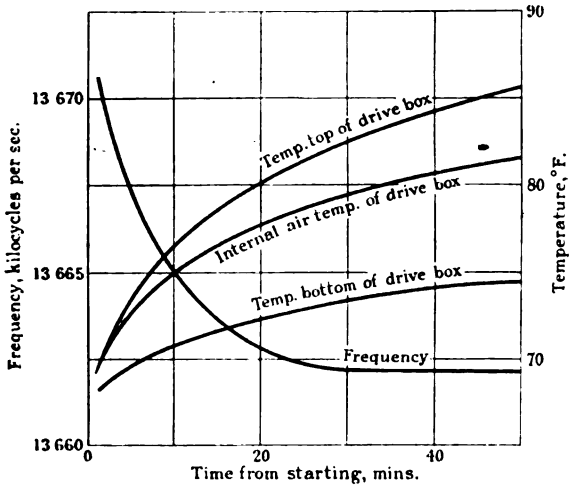

FIG. 5.—Initial frequency and temperature variation in beam transmitters.

Transmitter GLL (λ = 21·962 m), 22nd August, 1928.

Where special precautions are taken, as in the beam stations, the drift on starting-up may still be as much as 10 kilocycles per sec. in 20 000, or 1 in 10 000 if the transmitter is left unattended.

Some observations made by Mr. Pannett at the Dorchester station illustrate the effect of changes of

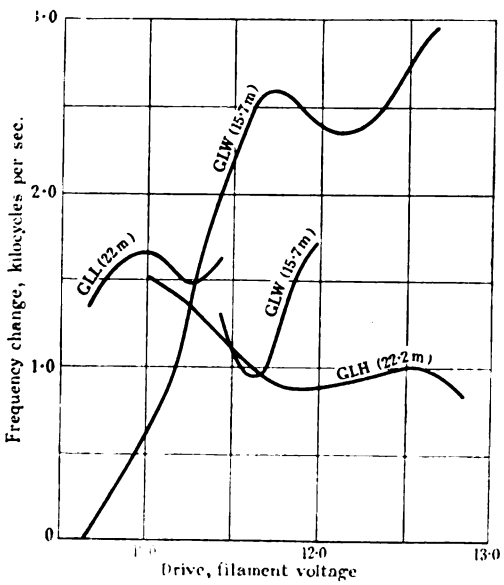

FIG. 6.—Frequency variation with drive filament voltage in beam transmitters.

filament and anode voltages and of temperature on the frequency of well-designed valve-driven transmitters.

Fig. 2 shows the effect of temperature during starting-up conditions, no adjustments being made during test to correct the frequency.

Fig. 3 shows the drift observed on a number of different frequencies. It is a straight line and shows that initial drift is roughly proportional to frequency.

Fig. 4 shows the effect of varying the anode voltage on three different transmitters.

Fig. 5 shows the initial drift on a transmitter compared with the rise of temperature measured in different parts of the master-oscillator chamber.

Fig. 6 shows the effect on the frequency of varying the filament voltage. The vagaries of these curves vary with individual valves.

Normally the engineer on watch makes the necessary adjustments during the starting-up period by checking against his standard wavemeter, and the outgoing fre-

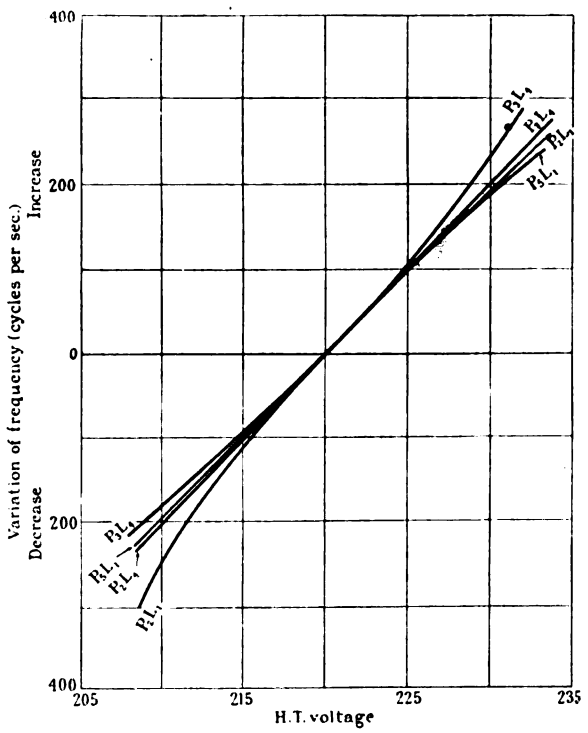

FIG. 7.—Variations of high tension on set No. 6 with various combinations of P610 and L610 valves in Franklin master-oscillator.

λ = 26·8 m (11 190 kilocycles).

quency is kept within the prescribed limits. Obviously the provision of a good wavemeter or frequency-measuring device is an absolute necessity in normal valve-driven stations, whether long- or short-wave, and the reputation of such stations depends largely on the reliability and precision of the wavemeter and the attention of the engineer-in charge.

In the new constant-frequency device two oscillator valves are used in combination and the principle of differential expansion between the insulating material constituting the former on which the coil is wound, and a brass rod carrying a small condenser of the vernier type, is made to compensate for changes in the L and C values of the oscillating circuit due to expansion or contraction. With this apparatus temperature control

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of the chamber in which the device is contained becomes unnecessary.

Adjustment of the compensator is made to correspond

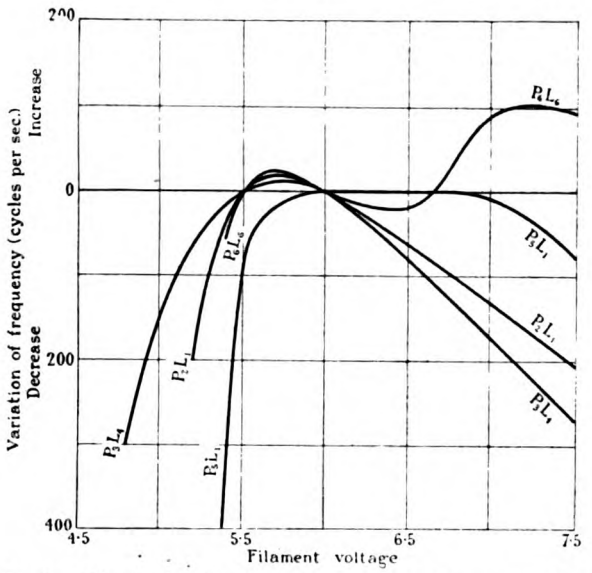

FIG. 8.—Filament voltage variations with various combinations of P610 and L610 valves in Franklin precision master-oscillator.

λ = 26·8 m (11 190 kilocycles).

to the centre of the scale, but it is found to be accurate over the whole scale of the standard instrument, which in one of its forms covers a 2 000-cycle range on a 15-m wave.

of 100 cycles in 11 190 000, or about 1 in 100 000. The four curves show the behaviour with various combinations of valves of the P610 and L610 types.

Fig. 8 shows the effect of varying the filament voltage. As might be expected, the lack of absolute uniformity among valve filaments shows up here, and the curves have divers shapes. A variation of 0·5 volt in the standard voltage applied to the 6-volt filament in no case makes a change greater than 100 cycles, or 1 in 100 000.

Fig. 9 shows the change in frequency of such an oscillator working on 15 050 kilocycles per sec., or approximately 20-m waves, when heated in a cabinet from the room temperature of 21° C. up to 46° C. and then allowed to cool again to a room temperature of 18° C.

During the first hour's heating the temperature of the cabinet rose rapidly to its final value of 46° C., yet there was no observed change in frequency of the oscillator as checked against a standard frequency meter. This is an important point as it shows that the frequency of the oscillator is not affected by sudden changes of temperature produced by draughts of air.

After about 5 hours' heating a small drift of approximately 250 cycles per sec. was noticed; this, however, did not change after a further 2 hours' heating. This change indicates that the oscillator is slightly overcompensated, and is noticeable only when the heat has penetrated right into the materials, especially into the insulating material, which has a low thermal conductivity.

After 7 hours the heaters were switched off and the oscillator cooled down slowly, overnight, in the cabinet. A measurement of the frequency made the next morning

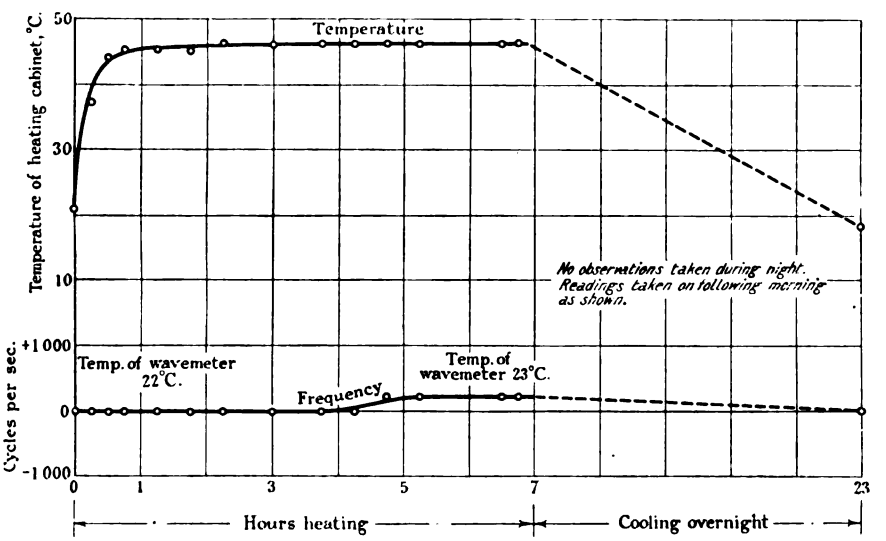

FIG. 9.—Drive for Rome–Sardinia service. Drift curve measured at output frequency of 15 050 kilocycles.

Observed drift for temperature-change of 25 deg. C. = 250 cycles.
Correction for wavemeter change due to 1 deg. C. temperature-rise = approximately — 100 cycles.
Drift ÷ 150 cycles with probable error of ± 250 cycles in reading wavemeter · therefore drift per deg. C. = 6 ± 10 cycles.

The curves in Fig. 7 show the variation of frequency, with variation of high-tension supply to the two master-oscillator valves. It will be noted that a variation of about 2½ per cent in h.t. voltage corresponds to a change indicated no change. During the heating the temperature of the frequency meter rose 1 deg. C., and a correction of minus 100 cycles per degree C. rise has to be made to any observed drift. As the scale of the

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frequency meter will only permit reading to 250-cycle divisions on its vernier, there is thus a probable error of ± 250 cycles in the actual observation of the oscillator drift measurement over 25 degrees. Thus the drift of the oscillator was 150 ± 250 cycles for a temperature-rise of 25 degrees C., or, in other words (+ 6 ± 10/15 050 000 cycles per degree C.), roughly about 1 in 1 000 000.

Broadcasting.

One of the outstanding events in broadcasting transmitter technique during the past year or so has been the tendency in Europe to adopt high-power transmitters. At international conferences a year and a half ago at Prague and last year at the Hague, arguments were put forward recommending the restriction of antenna power to about 100 kW.

The political and cultural value of broadcasting is so recognized among the nations that notwithstanding the splendid and successful efforts of the Union International de Radiophonie in adjusting to the best advantage the wavelengths, or should I say frequencies available for broadcasting, the total number of carrier frequencies allocated appears under present circumstances to be insufficient for the peculiar needs of our densely populated Continent with its great diversity of tongues. As a result, some of the nations are being obliged to make shift with a minimum number of exclusive frequencies and to increase the power and radius of their stations.

Notable high-power stations already working in Europe with C.C.I.R. aerial rating of 60 kW or more are Spanga, Rome, Oslo and Moscow, and Prague and Warsaw are in the course of construction.

High-power transmitting valves have made their commercial appearance in the design of broadcasting stations and perhaps some particulars of the new Marconi 100-kW valve, which I understand have not yet been published, will be of general interest. The valve, known as the C.A.T.10, which is rated at 100 kW input to the anodes on a telegraph load, stands about 3 ft. 6 in. high, its overall dimensions being 108 × 17 cm. The filament current is 225 amperes at 30 volts, and the anode voltage 10 000 to 15 000, the continuous anode dissipation being 50 kW. At 12 000 volts to the anode with zero grid bias the amplification factor is 45 and the impedance 3 500 ohms.

On long waves the normal operating conditions are:—

Telegraphy .. 8 amperes at 10 000–15 000 volts.
Telephony .. High-power modulation, 5 amperes at 12 000 volts.
Low-power modulation, 5 amperes at 12 000 to 15 000 volts.

A point of interest with regard to such valves is the problem of keeping down the glass temperature at the filament seal, the difficulty being that, owing to lack of space, such heavy filament currents as are required have to pass down into the tube through conductors of minimum cross-section. In this valve the problem has been solved by water-cooling the external portion of the valve filament leads.

In the lantern slide* the inlet and outlet tubes are clearly shown on the filament leads. Water flows down each lead to a point just above the seal, and returns again up the filament leg to the outlet tube.

Test figures for a broadcasting transmitter employing 8 of these valves, set for 80 per cent modulation and with an anode input of 405 kW at 12 000 volts, gave a carrier output to aerial of 120 kW, showing an efficiency of 29·6 per cent.

The artificial load employed for the test consisted of a morganite resistance of 465 ohms carrying 33·8 amperes.

A recent development in the design of high-power broadcast transmitting stations is the use of mercury rectifiers. Their high efficiency is an attractive feature in considering the maintenance costs of broadcasting stations, where the power to anodes runs into such figures as 400 to 600 kW. Three types of mercury rectifier are in use, viz. the mercury-vapour tube developed in America, the glass-bulb mercury-arc rectifier, and the Brown-Boveri type in steel cylinders.

In the transmitter test I have just mentioned, the Brown-Boveri type was used and showed an overall efficiency of approximately 95 per cent.

As a point of special interest I might mention the rectifier employed at the Moscow high-power broadcasting station, which I believe is the largest in use employing glass-bulb mercury-arc rectifiers. The photographs* show the construction of the rectifier panel, which consists of 12 three-phase rectifiers, each capable of handling about 30 kW with an efficiency of about 96 per cent. Power from the 6 500-volt 3-phase supply is taken to two transformers having star-connected outputs phased to give, in combination, a 6-phase supply to the rectifiers. The normal working load of this rectifier panel is 250 kW at 10 000 to 12 000 volts, with 8 bulbs in service.

In the design of high-power transmitters, whether for broadcasting or long-wave point-to-point communications, special precautions to prevent the radiation of harmonics become more and more necessary as the power is increased.

Apart from the use of astatic coils, capacity coupling between stages, and screening to prevent interaction between units, it becomes necessary on very long-wave working, as for the new 150-kW station at Ongar, to place the transmitter in a large room with copper screening to prevent the direct radiation of harmonics.

In large broadcasting stations it is becoming the practice to insert low-pass harmonic filters between transmitter output and aerial feeder circuit, and to screen all interconnections between the panels of the transmitter in earthed trunks to prevent jump-over, which would stultify the action of the filter.

Land-Line and Control Equipment for Broadcasting.

Referring to the land-line and control equipment field, the present trend of development in the design of speech amplifiers consists in giving careful attention to details and to individual parts of apparatus By so doing, a great improvement in overall frequency response has been effected.

* Not reproduced in the *Journal*.

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It will be readily understood that a moderate deviation from uniform response, which might be tolerated in individual pieces of apparatus taken separately, would become serious if such apparatus were used as part of a chain, and for this reason a much higher standard is required from individual parts of a chain than is required for the overall performance.

A study of the magnetic and electrical properties of the new series of nickel irons has led to improvement in the design and production of transformers to meet the stringent conditions already mentioned, and it has also made possible the use of much smaller and lighter transformers than hitherto.

By the use of carefully selected iron, built up in thin laminations, it has become possible to produce transformers having a uniform frequency response over quite a wide frequency spectrum, operative on radio frequencies, and it seems probable that this type of transformer may find many useful applications. Distortion arising in iron-core transformers will be largely eliminated as the conditions producing them are becoming understood and avoided.

By the use of new types of valves an economy in valve and battery use has been effected, accompanied by greater range of power output with distortionless working. Improved flexibility of control and switching has been obtained by the use of apparatus mounted on racks, and simple and effective means of mixing programme sources have been introduced.

Improved methods of indicating the instantaneous amplitude of the speech currents have made possible the more effective control of the depth of modulation at the transmitters, thus securing a greater average depth of modulation with less blasting caused by momentarily exceeding the distortionless modulation depth.

Conclusion.

In conclusion, and touching once again on the shortage of broadcasting wavelengths, and the tendency in Europe to employ high-power stations, it is interesting to note that there is a development taking place in the form of the establishment of local receiving stations, the function of which is to pick up the programmes of the high-power stations and re-broadcast them. This is done either on a low-power transmitter working on a common wave or on a very short wave below 10 m, or the received signals are distributed to the houses in the locality through a system of house-to-house wiring.

Short waves below 10 m have been found quite useful for local re-broadcasting. The range is small and unlikely to cause interference at a distance, and, subject to the obvious precaution of erecting the aerials on tall buildings clear of obstacles, good results are obtained. In this way first-quality broadcasting can be received in districts not normally served by the high-power stations.

The service of programme distribution through a house-to-house system is already in use in England and is considerably developed in several Continental countries, in particular in Russia, where for a small monthly subscription one can obtain the use of a good loud-speaker of the cone type and two programmes. It has been suggested that it would be cheaper to wire up a whole village and supply each house with a loud-speaker than to supply an equivalent number of receivers. Many of us might prefer not to be tied to the B.B.C. programmes, good as they are, and the fascination of listening to foreign stations is not to be despised. Nevertheless, I believe, if I may venture to prophesy, that we shall ultimately have wireless laid on in our houses, in the same way as the water, the gas, the electric light and the telephone. By then, however, it will no longer be wireless!

I am afraid that in attempting to review so wide a subject as radio communication and practice, I have only been able, as it were, to scratch upon the surface of things. I feel certain that I have omitted to mention important developments in some branch or other and I can only make the confession that specialization in the many branches of wireless design or exploitation is the order of the day and that it is becoming increasingly difficult to co-ordinate the various branches of investigation and development. Engineers are becoming specialized in the study of transmitting or receiving apparatus, long or short waves, direction-finding, valve manufacture, low-frequency apparatus, marine and aircraft apparatus, transmission and propagation phenomena, both with and without wires, even, as I have already mentioned, in the study of frequency allocations and the exploitation of channels in the ether, to say nothing of the erection of buildings, masts, plant and machinery. As a result, the erection of a station of importance is no longer a matter involving the selection for duty of a competent all-round engineer; it often involves the services of specialists in the final tuning-up and adjustment of a modern station.

Surely the lot of a radio engineer is varied, and his study, if he is interested in his profession, all-absorbing.

I should like to express my indebtedness to my colleagues of the Marconi Co., in particular to Messrs. Franklin and Wright, for assistance in collecting data for this address.

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BEAM ARRAYS AND TRANSMISSION LINES.*

By T. WALMSLEY, B.Sc.

(*Paper first received 4th March, and in final form 13th September*, 1930; *read before the* WIRELESS SECTION *3rd December*, 1930.)

SUMMARY.

The paper discusses the relative gains due to various combinations of half-wave radiators, and describes a type of array for which the author claims some economic advantages.

The causes of losses in transmission lines are considered and the conclusion is reached that these depend to a great extent upon the type of array used.

INTRODUCTION.

Although as early as 1899 Brown† had taken out a patent for a type of aerial which utilized the principle of interference, it was not until 1926 that the first commercially successful "beam" high-speed telegraph service was inaugurated at Bodmin, Cornwall. To Franklin belongs the honour of having designed the array system, and to Marconi the credit of having the vision and foresight to grasp the possibilities of short-wave beams. Earlier experimenters and scientists, notably Hertz, Stone, Blondel, Baun and Bellini, had prepared the ground, but their ideas were attended by little commercial success chiefly owing to the unsuitable wavelengths they employed or the difficulty of producing steady short-wave currents of sufficient amplitude. The development of the technique of short-wave generation enabled engineers to design transmitters capable of supplying a considerable output at steady frequencies. It was inevitable, therefore, that the ideas of early experimenters should finally materialize.

Since the establishment of the fact that short-wave beam telegraphy is a highly successful commercial proposition, many different types of beam arrays have been developed. The principle in all is the same: by means of wires correctly spaced, and currents correctly phased, an increase in field strength in the desired direction and a decrease in other directions are produced.

* Reprinted from *Journal I.E.E.*, 1931, vol. 69, p. 299.
† See Bibliography, (1).

TYPES OF ARRAY.

Various types of aerial are shown in Figs. 1, 2, 3 and 4. The Chireix-Mesny (French), Marconi, Koomans (Dutch) and Sterba (U.S.A.) all utilize wires excited as half-wave or multiple half-wave aerials. In the horizontal type of array, shown in Fig. 3, the horizontal wires are the radiators whilst the vertical wires act as feeders.

A type of array which has been used extensively in the U.S.A. and also in England for reception is shown in Fig. 5. As a receiving aerial, it has some advantages. Each vertical and horizontal member is a quarter of a wavelength long, an exception being made in the case of the two end horizontal members, which are each one-eighth of a wavelength long.

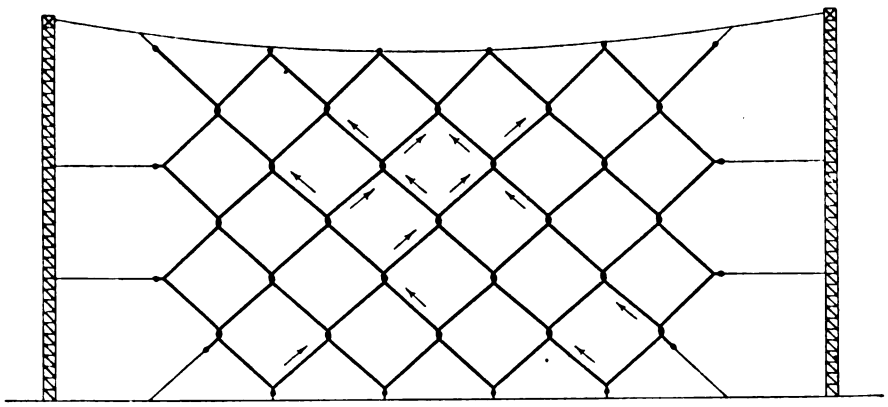

FIG. 1.—Chireix-Mesny short-wave aerial.

One of the purposes of this paper is to describe a new type of array developed by the author and illustrated diagrammatically in its elementary form in Fig. 6.

THEORETICAL PRINCIPLES.

To obtain continuity of reasoning, brief reference is here made to the theoretical principles underlying calculations of field strength due to arrays. Two assumptions are made in these calculations: (1) The distribution of current in the aerial wires is sinusoidal. (2) The earth is a perfect conductor.

With regard to the assumption of sinusoidal distribution of current, Wilmotte† has shown that, except when the lower end of the aerial is very close to the ground, the assumption is justified.

By taking from an aeroplane cinema records of signal-strength measurements due to a horizontal type of array erected at Farnborough and a Franklin-Marconi type of vertical array located at Grimsby, Wilmotte has also shown that, except at points very close to the earth, the

† See Bibliography, (3)

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measured and calculated polar diagrams—assuming a perfect earth—are very similar.

Accepting, therefore, as justifiable these two fundamental assumptions, the electric field intensity at a sufficiently great distance from a single unearthed half-wave vertical radiator is given by the expression developed by Levin and Young*:—

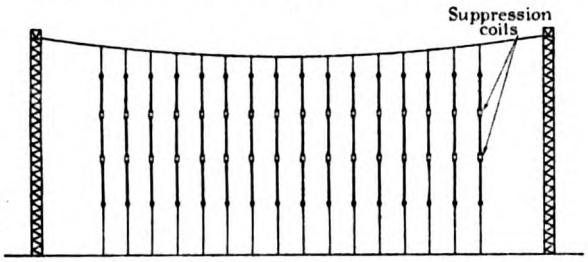

FIG. 2.—Marconi short-wave aerial.

$$\text{Intensity} = \frac{4a}{cr_0} \cdot \cos\frac{2\pi}{\lambda}(ct - r_0) \cdot \cos\left[\frac{\pi(\lambda + 4d)}{2\lambda}\cos\theta\right] \cdot \frac{\cos\left[(\pi/2)\cos\theta\right]}{\sin\theta} \quad . \quad (1)$$

where a = amplitude of the current at the centre of the aerial,
c = velocity of light,
r_0 = distance of the point considered from the aerial,
λ = wavelength = 2 × length of radiator,
d = distance of the bottom of the radiator above ground,
θ = angle between the vertical and line joining the distant point to the aerial (co-lateral angle).

To obtain the intensity of the electric field due to an array, the above expression should be multiplied by the array factor. Thus if the array consists of N similar wires spaced $\frac{1}{2}\lambda$ apart in a row, the amplitude of the electric field at a distant point will be

$$k\cos\left[\frac{\pi(\lambda + 4d)}{2\lambda}\cos\theta\right] \cdot \frac{\cos\left[(\pi/2)\cos\theta\right]}{\sin\theta} \cdot \frac{\sin\left[N(\pi/2)\sin\theta\sin\psi\right]}{\sin\left[(\pi/2)\sin\theta\sin\psi\right]} \quad . \quad (2)$$

where ψ is the angle between the vertical plane normal to the array and the vertical plane passing through the array and distant point.

This expression enables a polar diagram of field strength in any plane to be constructed.

Now Ballantine* has shown that the true criterion of effectiveness of a single vertical aerial is the ratio of the electric field intensity, E, in a desired direction to the square root of the power input, P, to the aerial.

Suppose, for example, that the desired direction of propagation from a single vertical wire used for broadcast purposes is horizontal. Then Fig. 7, which gives the vertical polar diagrams for a quarter- and a half-wave vertical aerial having their lower ends at ground-level, shows that since the area of the polar diagram due to the half-wave excitation is less than that due to the quarter-wave excitation, the half-wave aerial theoretically is more efficient than the quarter-wave. Eckersley† has shown that, as far as the ground wave is concerned, the actual improvement due to a half-wave vertical radiator over a quarter-wave vertical radiator is practically the same as the theoretical gain. The principle enunciated by Ballantine is also applicable to the question of effectiveness of different types of array. However, a difficulty at once arises:—Which is the desired direction of propagation for distant reception? In the horizontal plane the direction is obviously along

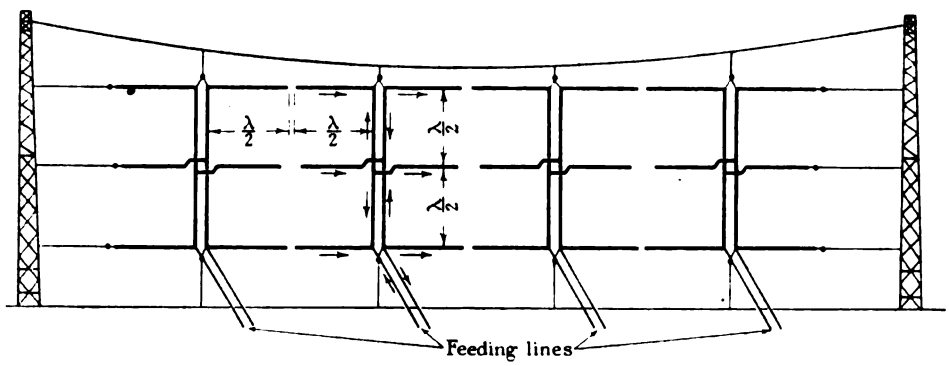

FIG. 3.—Koomans (Dutch) horizontal short-wave aerial.

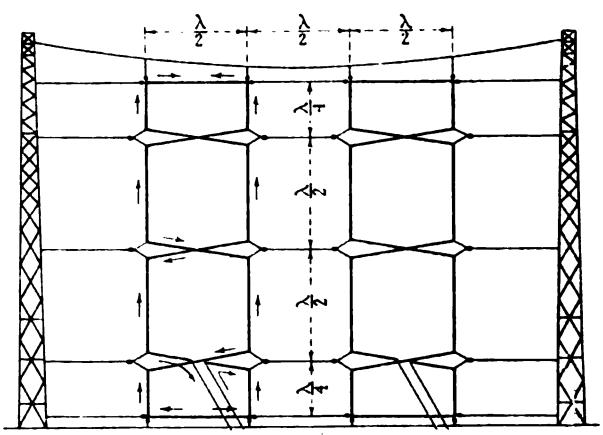

FIG. 4.—Sterba short-wave aerial.

* See Bibliography, 4).

* See Bibliography, (5). † *Ibid.*, (6).

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the great circle joining the transmitting and receiving stations. The most favourable direction in the vertical plane is not so easy to decide. Practical evidence, however, supports the contention that although the direction is not critical the best angle of projection is a few degrees from the horizontal. Recently the British Post Office Engineering Department and the American Telephone and Telegraph Co. conducted tests between America and England upon the received field strength in England due to two and three half-wave high arrays of the type shown in Fig. 4. The results showed that the improvement of these arrays over a single half-wave vertical aerial is such as might be expected on the assumption

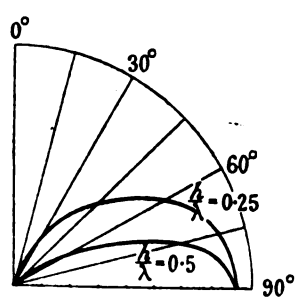

FIG. 7.—Polar diagrams in a vertical plane for single half- and quarter-wave vertical aerials having lower ends at ground-level.

that the earth is a perfect conductor and that the desired direction is horizontal, or almost so. Again, Meissner* has stated that the results of experiments between Germany and South America on 15 and 20 metres show that the best angle of projection is 0°, i.e. the most favourable radiation is approximately parallel to the surface of the earth. Meissner used in his experiment a horizontal aerial and a parabolic reflector.

* See Bibliography, (7).

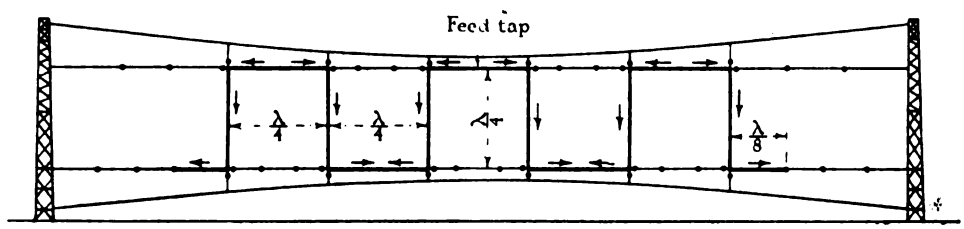

FIG. 5.—Bruce short-wave aerial.

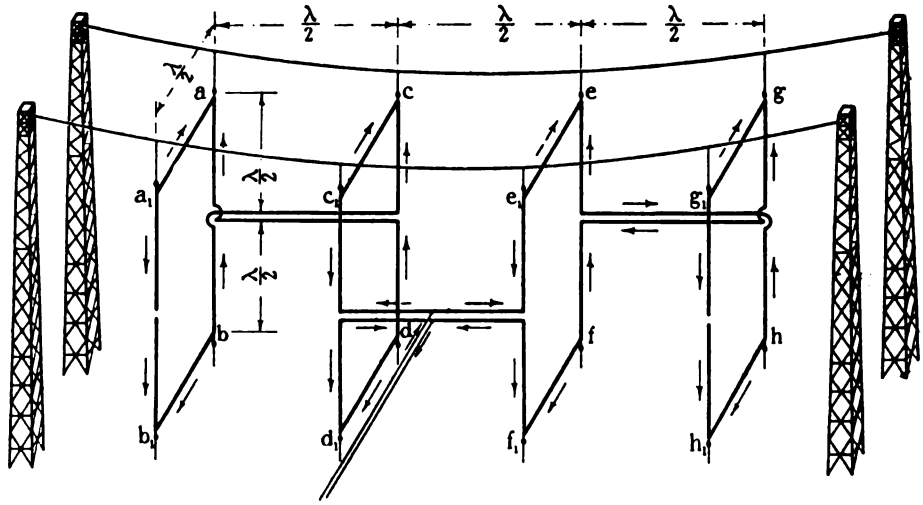

FIG. 6.—" T.W." short-wave aerial (single vertical unit).

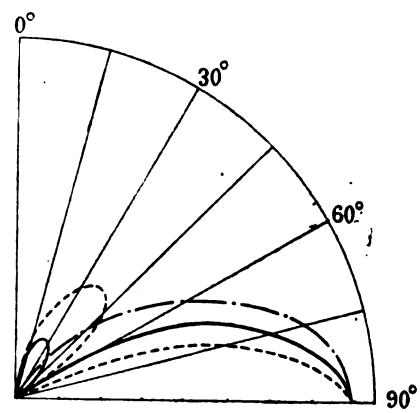

FIG. 8.—Polar diagram in a vertical plane for single half-wave vertical aerial having lower end at various distances above ground.

Lower end at ground-level — · — · —
Lower end ¼λ above ground ————
Lower end ½λ above ground - - - - -

It is, therefore, proposed to investigate the arrangement of vertical radiators that will give the best ratio $E/\sqrt{P}$ when E is taken in a horizontal direction, or,

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what amounts to the same thing, the best ratio of field strength in a horizontal direction due to an array of wires and due to a vertical half-wave aerial, the power input being constant.

The method of investigation is as follows:—Vertical radiators are arranged in three ways, (*a*) lined in a single row and spaced half a wave apart at the same height above ground, all currents being in phase; (*b*) arranged in two rows each similar to (*a*), one row being placed above the other in the same plane, all currents being in phase; (*c*) arranged in two curtains each similar to (*b*), the curtains being separated half a wavelength apart and having currents 180° out of phase; (*d*) arranged in four rows each similar to (*a*), placed in the same plane one above the other, all currents being in phase. These arrangements are illustrated in Fig. 13.

Methods of Computing Power Radiated from an Array.

Two methods of determining the ratio $E/\sqrt{P}$ are possible. In the first, an expression is obtained for the

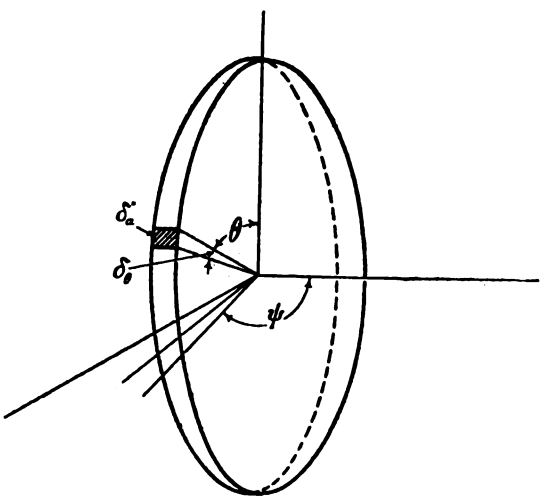

Fig. 9.—Polar radiation diagram.

value of the electric field E at a great distance from the array. By substituting this value in the Poynting vector $(c/4\pi)E^2$ and integrating, the total power radiated from the array may be obtained. The evaluation of the integral, however, presents considerable difficulty, and a computation of the total power radiated may therefore more readily be made by plotting the polar diagrams in several vertical planes.

Thus consider a distant point having polar co-ordinates (r, ψ, θ) (Fig. 9). If E is the value of the field strength at this point, the total energy per second radiated through a small area δa is proportional to $E^2r^2\delta\psi\delta\theta \sin\theta$. Thus, r being constant, the power radiated within the space enclosed by two vertical planes making an angle $\delta\psi$ with each other is proportional to $E^2\delta\psi\delta\theta \sin\theta$.

The quantity $\Sigma E^2\delta\theta \sin\theta$ may be obtained for a given value of ψ by tabulating values of $E^2 \sin\theta$ for several values of θ between 0° and 90°. By repeating this summation for various values of ψ, a curve may be drawn having $\Sigma E^2\delta\theta \sin\theta$ and ψ for ordinates and abscissæ respectively. Such a curve is shown in Fig. 10, which gives the case for 16 vertical synphased radiators each half a wave long spaced half a wavelength apart, and having their lower ends a quarter of a wavelength above ground. The area enclosed by the ordinates and the curve " a b c d " is a measure of power radiated by the array, and in order to compare with a half-wave aerial the rectangular figure " e f d o " equal in area to " a b c d o " has been drawn. Thus for equal power radiated, the improvement in field intensity in the direction $\psi = 0$, $\theta = 0$, is 10 log ("a o "/" e o") decibels. In the particular case considered this gain is 13·4 decibels.

The second method of calculating the radiation resistance of beam antennæ was proposed by Brillouin* and developed by Pistolkors.† In this method the e.m.f.'s

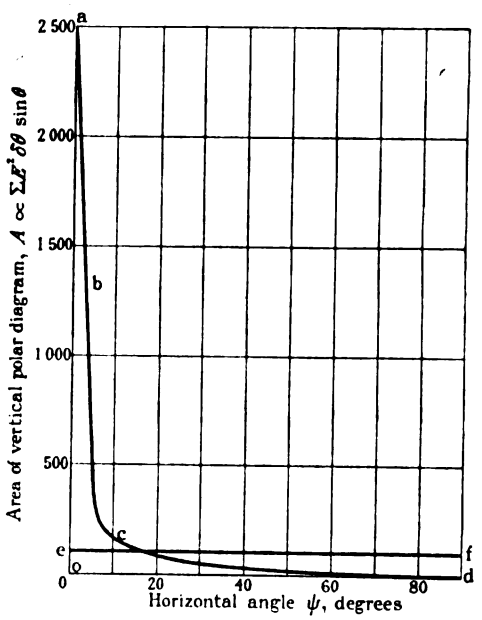

Fig. 10.—Comparison of power radiated by single vertical half-wave aerial and array of 16 half-wave wires spaced $\frac{1}{2}\lambda$ apart.

induced in the wires of an antenna by the current in the other wires are calculated. The total e.m.f. in each wire is the vector sum of the e.m.f.'s induced by the other wires and the e.m.f. associated with its own current.

The power, P, needed to suppress this total e.m.f. is regarded as the radiated power, and by dividing this quantity by the square of the current, taken at the middle of a half-wave element, the quantity termed " radiation resistance " is obtained.

Pistolkors has deduced expressions for the components of the power in one half-wave wire due to similar parallel wires spaced distances d apart horizontally and h apart vertically. The exact significance of d and h is shown in Fig. 11.

The second power component, P_0, depends on the current in the wire itself and is the limiting value of P_d when $d = 0$, $h = 0$. Thus the total power P radiated by one wire is

$$P = P_0 + P_d$$

* See Bibliography, (8). *bid.*, (9).

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and for both wires of the antenna system the radiated power is

$$2P = 2P_0 + 2P_d$$

If the currents I_0 in both wires are equal, the radiation resistance of the 2-wire systems is

$$2(P_0 + P_d)/I_0^2 = 2(R_0 + R_d)$$

Developing his arguments on the lines indicated, Pistolkors finally obtains expressions for the radiation resistance of a complex system of synphased half-wave radiators. For example, the system shown in Fig. 11, in which the image effect is taken into account, will have a radiation resistance

$$R_s = nR_0 + 2(n-1)R_1 + 2(n-2)R_2 + \ldots + R_{n-1}$$

where n is the number of radiators and

$$R_k = R(\tfrac{1}{2}k, 0) + R(\tfrac{1}{2}k, 2h_0 + \tfrac{1}{2}\lambda)$$

is the radiation resistance of an element due to the action of another radiator and its image. The term $R(\frac{1}{2}k, 0)$ denotes the resistance of the element due to a radiator spaced $\frac{1}{2}k$ horizontally and 0 vertically, whilst the image of this radiator—spaced $\frac{1}{2}k$ horizontally and

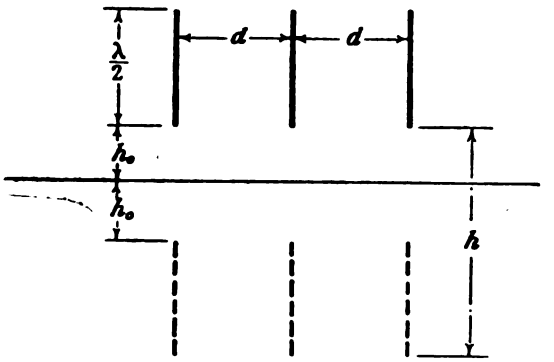

FIG. 11.—Half-wave radiators and images.

$(2h_0 + \frac{1}{2}\lambda)$ vertically—is responsible for the term $R(\frac{1}{2}k, 2h_0 + \frac{1}{2}\lambda)$.

By using the methods developed by Pistolkors for obtaining the radiation resistance of the elements of an array, and by utilizing the table given in the paper already quoted, the work of calculating the improvement of field strength in a desired direction due to an array is simplified considerably.

Thus let R_0 and R_N be respectively the radiation resistances of a single half-wave aerial and the average value for each half-wave element of an array consisting of N half-wave elements, and let r, which is assumed to be a constant value for all types of array, be the dead-loss resistance of each half-wave element. If the current amplitudes at the middle of the single aerial and each element of the array are I_0 and I_N respectively, then if equal power is supplied to each system

$$I_0^2(R_0 + r) = NI_N^2(R_N + r)$$

and the improvement in field strength in the desired direction

$$\frac{NI_N}{I_0} = \sqrt{\left[\frac{N(R_0 + r)}{(R_N + r)}\right]}$$

or array gain

$$= 10 \log_{10} \frac{N(R_0 + r)}{(R_N + r)} \text{ decibels.}$$

Using Pistolkors's* method of computing the radiation resistance, the gain of the 16-wire array previously mentioned has been calculated. The gain in this case when r is regarded as zero is the same as that obtained by the Poynting vector method, namely 13·4 decibels.

Effect on Radiation of Raising Radiators Above Earth.

Before comparing the values of E due to various arrangements of half-wave aerials, it is necessary to decide at what height above ground the lower end of the verticals should be placed, since this factor appreciably affects the radiation resistance of an array. In Fig. 12 curves have been plotted to show the effect of raising vertical half-wave aerials. It is apparent that in both of the cases considered, namely a single half-wave aerial and a

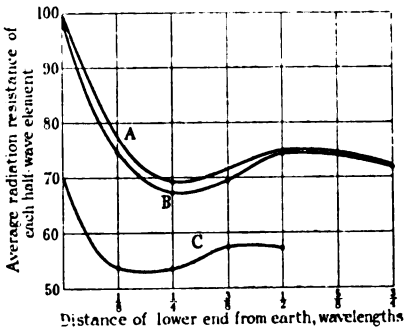

FIG. 12.—Variation of radiation resistance of vertical aerial with height above ground.

A.—One wire, calculated by "induced e.m.f." method.
B.—One wire, calculated by "Poynting vector" method.
C.—Seven similar $\frac{1}{2}\lambda$ vertical radiators spaced $\frac{1}{2}\lambda$ apart, calculated by Pistolkors's method.
NOTE.—In all cases perfect earth is assumed.

7-half-wave aerial, no reduction in radiation resistance is obtained by raising the lower end of the aerial more than a quarter-wavelength above earth. This value is therefore taken in all subsequent calculations.

It seems somewhat paradoxical that reduction of radiation resistance in this manner is actually productive of increased signal strength. This seeming paradox, however, is at once explained when it is realized that as the aerial is raised the radiation in directions making angles with the horizontal is at first reduced, and thus for a constant amount of power input to the array the value of E increases. This increase, however, is not maintained when the aerial is still further raised, owing to the production of secondary loops of radiation in vertical planes. Fig. 8 shows this effect.

Effect on Radiation of Grouping Radiators in Various Formations.

Using Pistolkors's methods, the gain over a half-wave radiator due to various groupings of radiators has been calculated and the curves of Fig. 13 have been plotted. The value of r in these calculations has been taken as

* See Bibliography, (9).

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25 ohms. From these curves, the following conclusions can be drawn:—

(1) The best ratio of $E/\sqrt{P}$ is obtained by arranging a number of half-wave wires in a single row at equal distances from the ground. For example, 4 wires arranged in groups A, B, C and D (Fig. 13) show the following gains over a half-wave aerial: 6·85, 5·8, 4·1 and 4·5 decibels respectively; 8 wires arranged in a similar manner give 10·0, 9·0, 6·9 and 8·7 decibels gain respectively, whilst 16 wires result in gains of 13·0, 11·9, 10·3 and 11·7 decibels respectively.

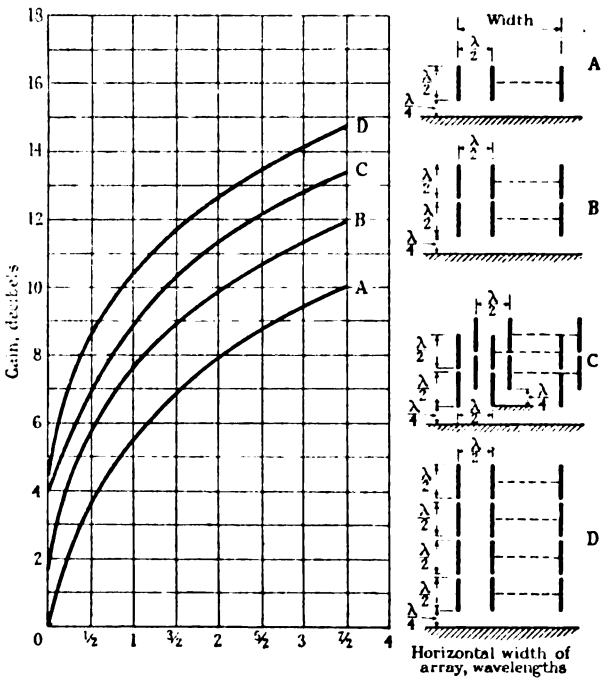

FIG. 13.—Gain of various types of array over half-wave vertical aerial raised $\frac{1}{4}\lambda$ above the ground (theoretical values).

A.—Single $\frac{1}{2}\lambda$ verticals spaced $\frac{1}{2}\lambda$ apart in one plane, all wires $\frac{1}{4}\lambda$ above ground, currents in phase.
B.—Double $\frac{1}{2}\lambda$ verticals spaced horizontally $\frac{1}{2}\lambda$ apart, all lower wires $\frac{1}{4}\lambda$ above ground, currents in phase.
C.—Two lines as in B, lines separated by $\frac{1}{2}\lambda$, all currents in exciter line in phase, all currents in back line in phase but lagging 180° between front and back.
D.—Four $\frac{1}{2}\lambda$ verticals spaced horizontally $\frac{1}{2}\lambda$ apart, lower wires $\frac{1}{4}\lambda$ above ground, all currents in phase.

(2) For a given span of array, the gain above a half-wave aerial is the greatest in the case of the arrangement D. For example, the gains for a span of $3\frac{1}{2}$ wavelengths between extreme wires of the array are 10·0, 11·9, 13·4 and 14·7 decibels for arrangements A, B, C and D respectively. The difference in gain of the various arrays which is shown by these values is maintained approximately without change as the span increases. Thus both span and height are important factors in obtaining an array having a large gain.

ECONOMICAL SIZE OF ARRAY.

Economical considerations, however, limit both span and height. To minimize the uneconomical traffic periods due to fading, it is customary to install arrays for two, three, or even four wavelengths. This fact, combined with the policy of centralization of services, results in a very large number of arrays being required on the same site. A large site is required to accommodate the arrays, which must necessarily have different orientations and considerable space separation. Thus the cost of both land and arrays is increased. Moreover, any increase in spans and in distance from the central transmitting station necessitates an increase in the lengths of feeding lines, with corresponding increase in capital expenditure and line energy losses. Finally, as pointed out by Friis, lateral deflection of beams is frequently experienced, thus reducing the usefulness of wide arrays that produce very narrow beams. With regard to an increase in the height of an array, any extension vertically necessitates increased height of towers. Some increase can be made at reasonable cost, but since the price of towers rises almost exponentially with height an economical limit is quickly reached. For example, in the case of an array consisting of exciting and reflecting panels for a wavelength of 16 metres a convenient span is 8 wavelengths, which, allowing for attachments and clearance between supporting structures and end radiators, would necessitate two towers separated by 500 ft.

An array having two rows of vertical radiators one above the other would require towers $2\frac{1}{4}$ wavelengths or 120 ft. in height. This figure is taken from an actual

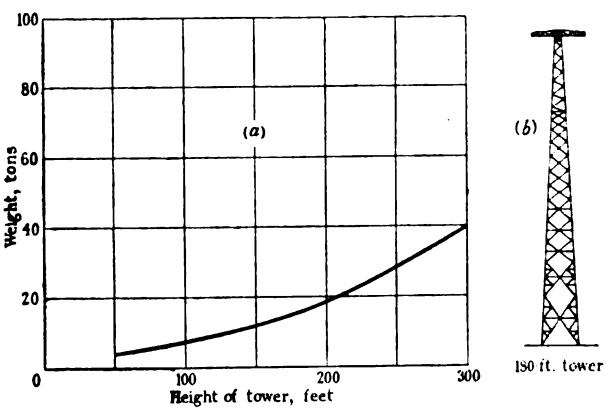

FIG. 14.—Weight of self-supporting towers. Horizontal top pull 4 tons, torque 26 ton-feet, top downthrust $1\frac{1}{2}$ tons.

design by the author and allows one-quarter of a wavelength clearance from the ground, three-quarters of a wavelength for dip of the supporting cables, and one-quarter of a wavelength for insulators and attachments. Again, the towers would be required to resist a total top pull of 4 tons, viz. 2 tons for the exciter panels and 2 tons for the reflectors. An array having 4 vertical rows of half-wave radiators would require twice the dip allowed for the 2-row array to ensure a 4-ton top pull. The height of the towers required would thus be about 4 wavelengths, i.e. 210 ft. Reference to Fig. 14(*a*), which gives the weight of steel required for lattice towers (designed by the author) of the type illustrated in Fig. 14(*b*), shows that the weights of steel for 120-ft. and 210-ft. towers are $9\frac{1}{2}$ tons and 20 tons respectively. Assuming an approximate cost for fabricated steel of £40 per ton inclusive of foundations and erection, or a total of £760

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and £1 600 respectively for the two towers for each array, and estimating the cost of the arrays complete with insulators as £350 and £650 respectively, the grand totals are £1 110 and £2 250 respectively. The field-strength gain of the four vertical half-wave aerial over the two vertical half-wave aerial has previously been shown to be 2·8 decibels, and this gain might be well worth the extra expenditure involved. As, however, arrangement C has only 1·3 decibels less gain than arrangement D (Fig. 13), it is proposed to examine the relative costs of the two systems.

Reflector Curtains.

Hitherto the effect of reflector wires has been disregarded, and the curves of Fig. 13 refer to the exciter portion of the various systems only. For a true comparison of the two systems C and D, consideration must be given to reflectors. Taking the case of a reflector curtain consisting of a number of half-wave vertical elements in the same plane, it is at once obvious that

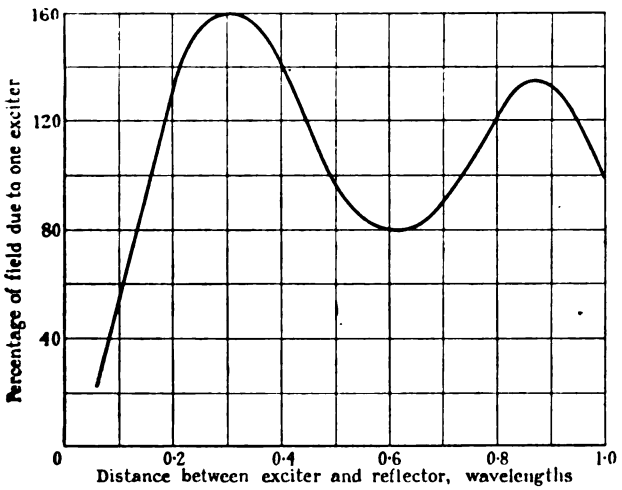

Fig. 15.—Field strength in forward direction due to one exciter and one reflector spaced various distances apart.

system C may not lend itself to so efficient a reflector curtain as system D, since the front wires of the exciter curtain must be a half-wave further away from the reflector curtain than the back wires and only one reflector curtain is provided for the two exciter curtains. Usually the spacing between exciter and reflector wires is approximately one quarter of a wavelength. Three-quarter-wave spacings, however, are sometimes used, as, for example, in one of the Marconi-Franklin beam systems at Bodmin.

The curve of Fig. 15 has been prepared from a theoretical investigation carried out by Sammer (reported by Gothe*) into the effect of a single vertical half-wave tuned reflector spaced at various distances behind a single vertical half-wave exciter wire. The combined effect upon the field strength in a forward direction is shown to be greatest when the spacing between exciter and reflector is approximately 0·32λ. By increasing the spacing to 0·87λ a second peak of field strength is obtained. These results are in substantial agreement with those of Englund and Crawford,* who concluded from their experiments with Hertzian doublets that spacings for maximum peaks were very nearly 0·33 and 0·83 of a wavelength. Regarding the field due to a single exciter as unity, the intensities due to the combination of two wires are 1·6 and 1·35 times as great respectively in the two cases. It would thus appear that a considerable reduction in the forward field strength is incurred by spacing the reflector 0·87 instead of 0·32 of a wavelength behind the exciter. When the number of exciter and reflector wires is increased, however, this reduction is not so great. For example, tests on a 16-metre (approx.) Marconi-Franklin array having 48 vertical lines of radiators and 104 vertical lines of reflector wires spaced three-quarters of a wavelength behind the exciter curtain showed that the ratio of maximum front to maximum back radiation, measured at ground-level locally, was better than the ratio obtained with a 26-metre (approx.) array of similar type consisting of 32 vertical lines of exciters and 64 lines of reflectors, spaced one quarter of a wavelength behind the exciter curtain.

In order to ascertain the effectiveness of one radiation-coupled reflector curtain placed behind two curtains of exciter wires spaced half a wavelength apart, the author made tests on the type of array shown in Fig. 16. Three curtains were erected, two consisting of 6 and one con-

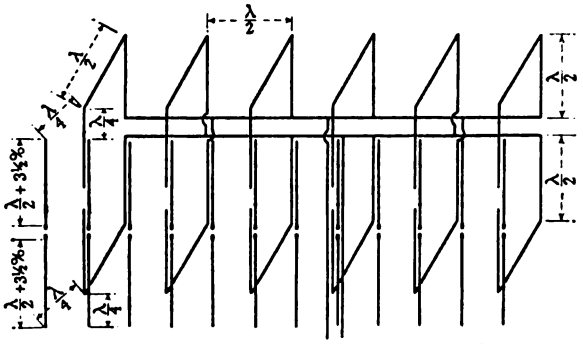

Fig. 16.—Double-row vertical array with reflector.

sisting of 12 vertical lines of half-wave radiators. All were two half-waves high, i.e. 16 metres approximately. The first two curtains, which were spaced half a wavelength apart, were fed directly from a transmission line; the third curtain was spaced three-quarters and one-quarter of a wavelength approximately behind the other two curtains respectively, and was "radiation-coupled." Adjustments were made in the lengths of the reflector wires and the spacing from the exciter curtains until the maximum forward field strength was obtained. The best effect was obtained when the wires of the reflector curtain were made 3½ per cent longer than half a wavelength and were spaced one-quarter of a wavelength apart and one-quarter of a wavelength away from the nearest exciter radiators. In these circumstances the horizontal polar diagram shown in Fig. 17 was obtained, from which it is seen that the ratio of front to back field-strength measured locally was 4 : 1. A later result with a slightly different arrangement gave a ratio of 12 : 1. This result is as good as the best obtained with

* See Bibliography, (10).

* See Bibliography, (11).

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ordinary single-curtain exciters and reflectors. Moreover, complete reflection would only increase the forward field by about 2 per cent. These facts therefore justify the contention that the addition of a reflector curtain will not affect the relative gain of the D and C types of array.

To support the C type, towers similar to that shown in Fig. 14(*b*) are required. The top cross-arm, however, must have a span of three-quarters of a wavelength instead of one-quarter. Very little extra steel is required on this account. Allowing a generous addition for the cost of increased span of arms and for additional attachments required for the support of the aerial, the estimated total cost of an array and its supporting masts 120 ft. high is £1 300.

Thus, to gain 1·3 decibels improvement over the C type of array, an additional expenditure of £950 per array would be involved. On the other hand, a C type of array would give a gain of 1·5 decibels over the B type for an additional cost of £190.

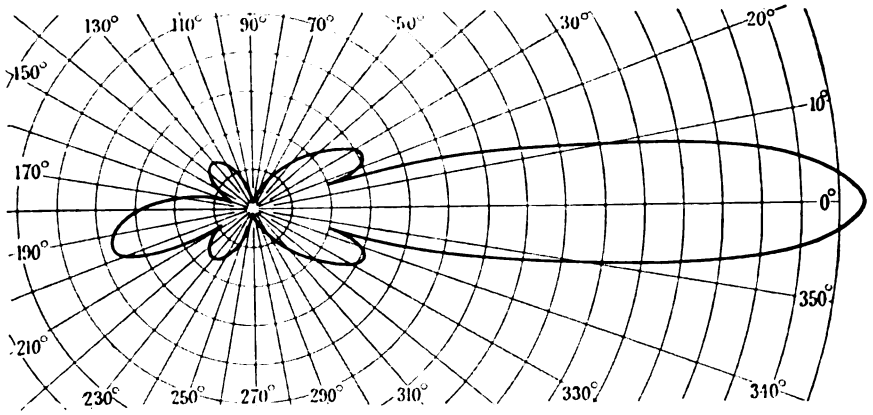

FIG. 17.—Measured horizontal polar diagram of array shown in Fig. 16.

of the latter type is shown diagrammatically in Fig. 18. Each unit is "current-fed," i.e. the attachment to the transmission lines is made at the centre of a half-wave element at the position of minimum impedance.

The phase of the current in the various half-wave sections is shown conventionally by arrows. Thus in Fig. 6 all the lines of radiators "ab," "cd," "ef," "gh," have currents in phase with each other, whilst all the lines "a_1b_1," "c_1d_1," "e_1f_1," "g_1h_1," have also currents in phase with each other but 180° out of phase with those in the first lines. Antinodes of current and nodes of potential occur at the middle of each half-wave radiator. The side radiation from the elements "aa_1," "bb_1," "cc_1," "dd" might at first sight be thought to be considerable. However, the spacing of these elements and the phase of their currents is such as to produce a comparatively small amount of side-radiated energy. An approximate calculation of the energy radiated from these members shows that the resultant loss per unit is about 16 per cent of the total energy supplied to each. With the horizontal type of array (Fig. 18) in which several units are used, the wires "bb_1," "dd_1," etc., of each unit are in close proximity to similar wires of the adjacent unit—except at the two ends of the array. The currents in these adjacent wires are 180° out of phase and thus side radiation is almost entirely cancelled. In the case of the assembly of vertical units such mutual cancellation is not experienced.

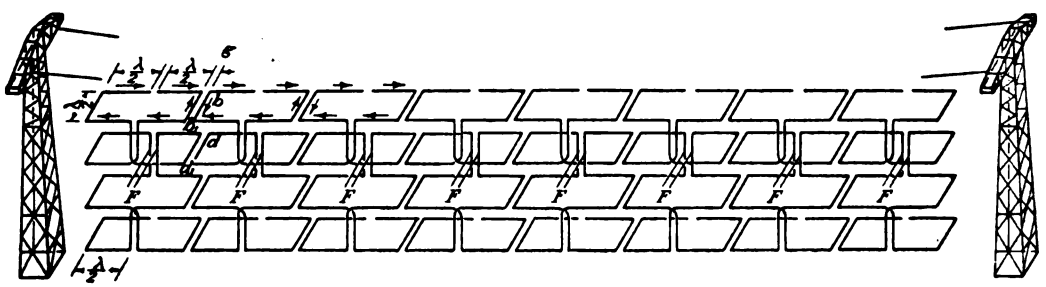

FIG. 18.—"T.W." horizontal-type array (diagrammatic).
Feeding-points shown at F.

"T.W." VERTICAL TYPE OF ARRAY.

Considerations such as these led the author, on behalf of the Post Office Engineering Department, to devise and build arrays consisting of a number of units of the type shown in Fig. 6. In this figure the reflector curtain has been omitted for the sake of clearness.

The units may be used in groups either of vertical radiators, as shown in Fig. 6, or—by rotating the whole system through 90°—of horizontal radiators. An array

Reduction of Unwanted Radiation.—Cancellation can to

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a large extent be obtained, however, by means of wires, insulated at their ends and approximately half a wavelength long, stretched parallel to the offending elements. According to Sammer,* when the space separation between wires is 1/16th (0·0625) of a wavelength the phase separation between the currents in the two wires

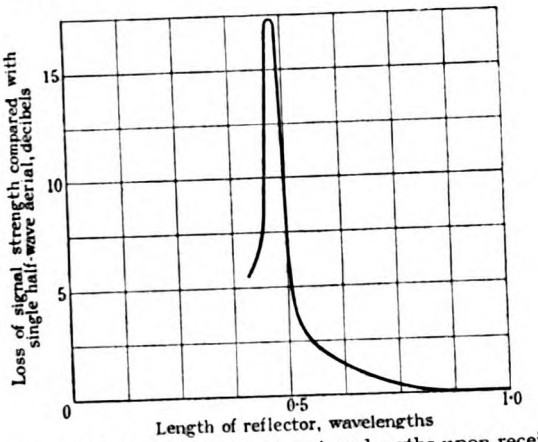

FIG. 19.—Effect of reflectors of various lengths upon received signals.

Wavelength = 16·10 metres.
Length of excited wire = 8·05 metres.
Distance apart of exciting and reflecting wires = 1 foot.

is 170°, and the currents have approximately equal amplitudes. Under these conditions the energy radiated from the dual combination is calculated to be about 17 per cent of the energy radiated from the single radiator.

Tests made for the author by Mr. Mumford confirmed qualitatively the fact that radiation from an excited wire can be greatly reduced by means of currents induced in a parallel wire. Two vertical wires were erected; the exciter wire was half a wavelength high and was connected to a measuring set through a 3-ft. length of horizontal wire, while the reflector wire was 0·72 wavelength long originally and was completely insulated.

A portable oscillator feeding a short vertical aerial was placed a few wavelengths away from the receiving aerial, the transmitting aerial and the two wires of the receiving aerial being in the same plane. Measurements were made of the received signal strength for various values of the length of the reflector wire. The plotted results are given in Fig. 19, which illustrates the critical nature of the suppression effect.

Tests with a T.W. vertical-type array were also made to ascertain the effect on the efficiency of the array of insulated wires a little less than half a wavelength long. It was found that the ratio of field strength—measured several wavelengths away—to array input improved appreciably when the suppression wires were used.

Open transmission lines feed both types of T.W. arrays. Well-known principles of matching correctly the termination and junctions of transmission lines have been tried and tests made with combinations of coils and condensers, as suggested by Chireix,* Moser† and other experimenters. However, as these are difficult to adjust, and, unless well protected, variable in behaviour under different climatic conditions, a more practical solution is offered by designing transmission and feeding lines to have correct surge impedances. The multi-wire line described in a later part of the paper lends itself admirably to this purpose, as variations in surge impedances from 150 to 600 ohms can readily be obtained. It was pointed out by Beverage, Rice and Kellogg‡ in 1923 that, when standing waves occur along a transmission line, maximum and minimum peaks of voltage are produced at distances separated by multiples of a quarter of a wavelength. They also showed that the surge impedance of the line is the geometric mean of the apparent impedances at the points of adjacent quarter-wave peaks. By applying these principles and choosing the best positions for the junctions of feeders and transmission lines, reflections can be greatly reduced. The question of transmission lines is considered in greater detail in a later section.

Two practical difficulties are encountered in fixing suitable matching devices to some types of array. These are due firstly to the height above the ground of the feed position, and secondly to the wide difference in the impedances of the array and the feed line. For example,

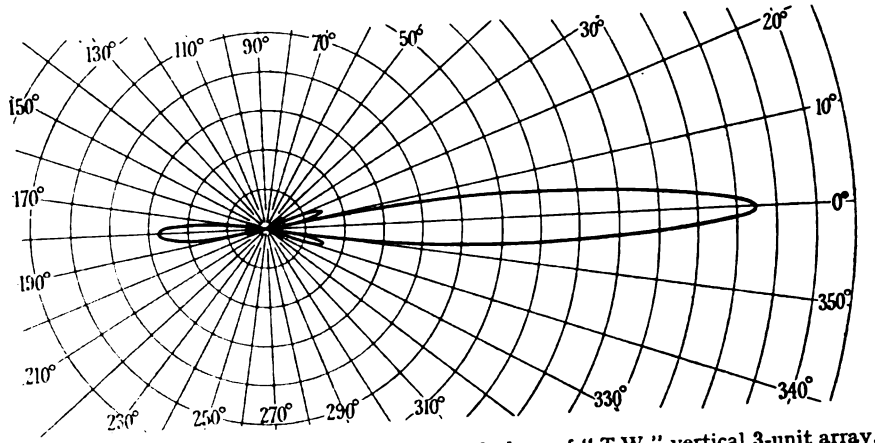

FIG. 20.—Measured polar diagram in horizontal plane of "T.W." vertical 3-unit array.

Span between end radiators = 5½λ.

* See Bibliography, (10).

* See Bibliography, (12). † *Ibid.*, (13). ‡ *Ibid.*, (16).

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the feed position of the Chireix-Mesny array (Fig. 1) is half-way up the aerial, and in consequence the difficulties of fixing and adjusting the matching devices at this height are considerable. It is, of course, possible to

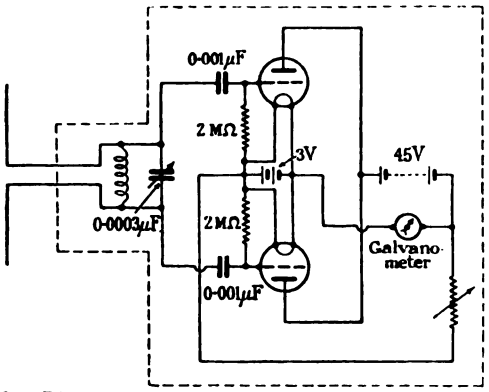

FIG. 21.—Diagram of connections of portable field-measuring set.

bring the feeding lines nearer to the ground and fix the matching devices at the lower level, but to obtain the maximum transference of energy from the feeding lines to

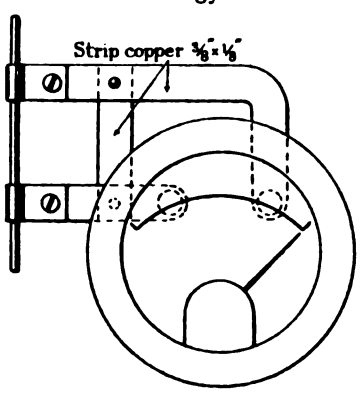

FIG. 22.—Shunted radio-frequency ammeter.

the array the matching devices should be fixed close to the array unless the impedance of the latter is approximately the same as that of the feeding lines. The horizontal type of array (Fig. 3) in which several half-wave radiators are connected in parallel presents an impedance to the feed lines that might differ considerably from that of the latter. Owing to the series arrangement of radiators this disadvantage is not shared by either the Sterba or the T.W. array, which, when correctly adjusted, have an impedance approximately equal to that of the open two-wire transmission lines.

Tests and Polar Diagrams.—Comparison between a half-wave vertical receiving aerial and a T.W. 16-metre vertical-type array consisting of four units, i.e. having a span between extreme wires equal to 7½ wavelengths, disclosed a gain of 20 decibels. Allowing 3 decibels for the reflector, the maximum theoretical gain was 19·4 decibels.

In these tests a small vertical aerial excited by a portable oscillator was located many wavelengths away from the two receiving aerials, which themselves were separated by several wavelengths to avoid interference effects. The method of measuring the received signal strength was the same as that described by Friis and Bruce.* The discrepancy between calculated and measured values is well within the limits of experimental error for this class of test.

In Fig. 20 the measured polar diagram in a horizontal plane of a 3-unit T.W. transmitting array, working at a 20·78-metres wavelength, has been plotted. Owing to limitations of space, readings were taken at a radius of 750 ft. only, and in consequence a slightly broader main beam was obtained than would have been measured at a greater distance. Corrections have therefore been made, and the resultant diagram gives the comparative values of the field strength at infinite distance from the array. The absence of small protuberances is explained by the fact that they were too small to be measured accurately.

The field-strength measurements were made by means of a push-pull valve voltmeter, coupled to a small dipole aerial (Fig. 21). A crystal-controlled oscillator fed the array, supplying it with about 5 kW of power. Measurements of the current in the transmission lines were made by clipping copper connections across a definite length of line by means of the clips shown in Fig. 22.

* See Bibliography, (14).

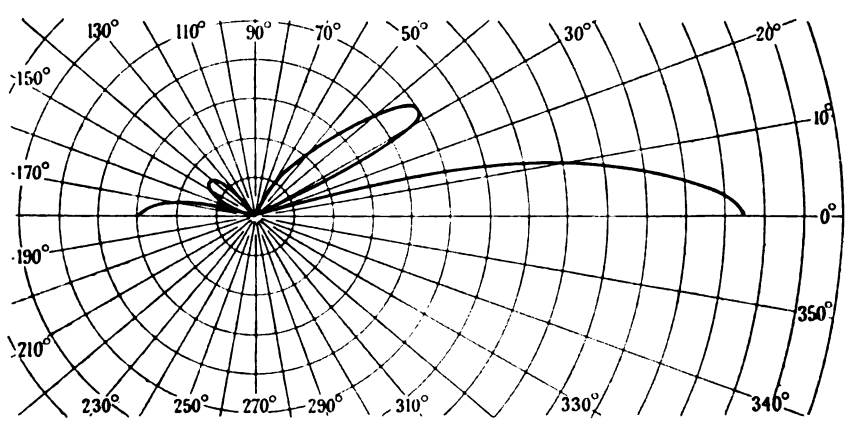

FIG. 23.—Calculated polar diagram of "T.W." vertical-type array, taken in vertical plane normal to array.
Height of array = two ½λ; lower radiator ½λ above ground.

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It was not possible to measure the polar diagram in a vertical plane normal to the array, but, regarding the earth as a perfect conductor, the theoretical values have been plotted in Fig. 23. To calculate the comparative values of field strength the expression (2) for the field strength due to a single row of vertical radiators was utilized. The field due to a single curtain was first obtained by adding vectorially the fields due to the lower vertical radiators and their images to that due to the upper radiators and their images. Thus, ψ being zero and d having the value $\frac{1}{4}\lambda$ for the lower radiators and $\frac{3}{4}\lambda$ for the upper, the strength of the field at a distant point is obtained by substituting for

$$\cos\left[\frac{\pi(\lambda + 4d)}{2\lambda}\cos\theta\right]$$

in formula (2) the expression

$$2\cos\left[(3\pi/2)\cos\theta\right].\cos\left[(\pi/2)\cos\theta\right].$$

The vector sum of the fields due to each of the three curtains gives the value of the field due to the whole array. Now the field strengths due to the second (exciter) and third (reflector) curtain differ in phase from the field due to the first curtain by $\pi(1 - \sin\theta)$ and $1\frac{1}{2}\pi(1 - \sin\theta)$ respectively, assuming at the reflector a phase lag of 180° between the incident and re-radiated wave.

Denoting these phase angles by A and B respectively, the array factor due to vector addition becomes

$$\sqrt{\{2 + \gamma^2 + 2(\cos A + \gamma\cos B) + 2\gamma[\cos(A - B)]\}} \quad (3)$$

in which γ is the ratio of the sum of the currents in the reflector wires to the sum of the currents in the wires of each exciter curtain. In the particular case considered γ was 1·25, and by using this value in expression (3) the diagram of Fig. 23 was obtained.

"T.W." Horizontal Type of Array.

Considerable attention has been given recently to arrays consisting of groups of horizontal radiators. In Germany and Holland particularly this type has gained favour.

The behaviour of a T.W. horizontal type of array consisting of eight units is therefore of interest. The vertical polar diagram of this array, calculated by methods similar to those adopted in the case of the vertical array, is shown in Fig. 24.

In a vertical plane normal to the array, the field at infinite distance due to the four rows of horizontal radiators in the front curtain and their corresponding images has an amplitude proportional to

$$\frac{\sin^2(2\pi\cos\theta)}{\sin(\frac{1}{2}\pi\cos\theta)}.$$

As the reflector curtain in the particular horizontal type of array under consideration has only half the number of wires of the two exciter curtains taken in conjunction, the value of γ in expression (3) is less than in the case of the vertical type. Actually the value was unity, i.e. the measured current in each reflector wire was approximately equal to the current in each exciter wire.

Reference to Fig. 24 shows that the direction of maximum field makes an angle of 10° with the horizontal.

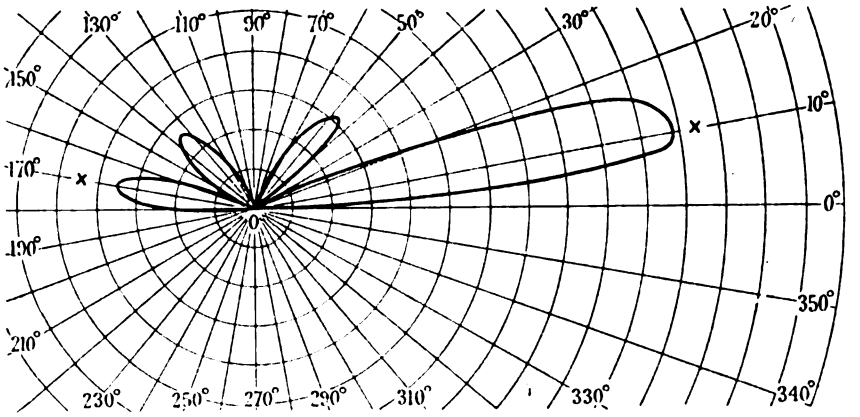

Fig. 24.—Calculated polar diagram of "T.W." horizontal-type array, taken in vertical plane normal to array.
Currents in all exciting and reflecting elements considered to be equal; lower radiator $\frac{1}{4}\lambda$ above ground.

Further, there is no horizontal radiation. Thus, if theory and practice agree, it should not be possible to make local measurements of field strength at ground-level. This is actually the case, and in consequence a measured horizontal polar diagram cannot be plotted. However, measurements made at long distances from the array show that the theoretical expectation for this type of array can be realized.

Transmission Lines.

General.—Scarcely less important than the question of suitable types of array is the problem of suitable transmission lines. The contention is frequently made that, owing to their low energy losses at radio frequencies, concentric metallic tube transmission lines are economically essential. The Marconi beam telegraph stations in England and abroad, the French St. Assise station, and the German Nauen station all use concentric tube lines. For reception purposes, the "pick-up" with open transmission lines is considered by many radio engineers to be so serious as to prevent their use even on short lines. It is proposed in this section to contest this view

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by showing how losses on open transmission lines may be reduced to such an extent as to render them successful competitors to the concentric tube lines. Formulæ are first obtained to enable the losses in transmission lines to be determined, and by applying these formulæ the losses in several open lines—connected to various types of array—are calculated. An analysis of these losses is then made, and as a result the opinion is advanced that the type of array used influences the amount of losses. A special type of multi-wire line is finally described.

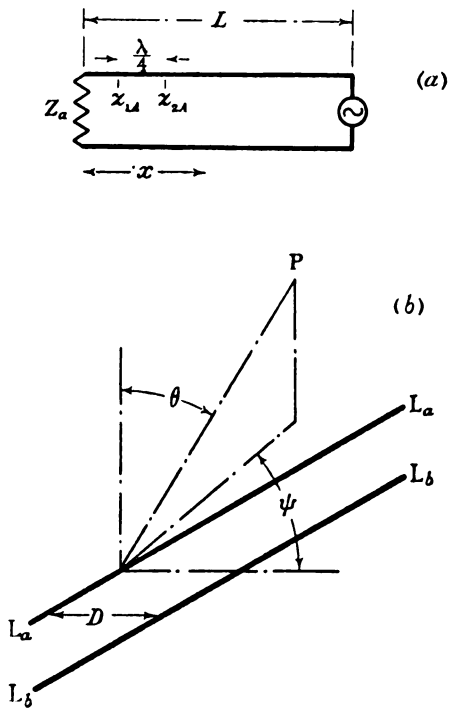

FIG. 25.—Transmission-line diagrams.

Theoretical Principles.—The impedance Z at any point in a transmission line at a distance X from the load may be represented by the expression—

$$Z = Z_0 \cdot \frac{(Z_0 + Z_a)e^{ax}e^{jBx} + (Z_a - Z_0)e^{-ax}e^{-jBx}}{(Z_0 + Z_a)e^{ax}e^{jBx} - (Z_a - Z_0)e^{-ax}e^{-jBx}} \quad (4)$$

where Z_a = terminal impedance,
Z_0 = surge impedance of the line,
a = attenuation constant,
B = velocity constant $= 2\pi/\lambda$ at radio frequencies,
$j = \sqrt{-1}$.

Using this expression, and utilizing the fact that the terms e^{jBx} and e^{-jBx} indicate two vectors rotating in opposite directions, Nancarrow* has shown that maximum and minimum values of impedance, voltage and current occur at quarter-wave intervals along a transmission line when Z_a is not identical with Z_0.

Further, he shows that the impedance at the points of maxima and minima is purely resistive. Beverage, Rice and Kellogg pointed out the same facts in 1923.† These facts suggest a method of measuring the electrical losses in lines feeding aerials.

* See Bibliography, (15). † *Ibid.*, (16).

The method of procedure is to obtain values (by calculation) for Z both at the transmitter and at the aerial end of the transmission line. The current $I_{min.}$ is measured at each of these points, and since $Z_{max.}$ is purely resistive the input and output values of the power I^2R can be ascertained.

One difficulty immediately arises: although Z_0 at high radio-frequencies is resistive, Z_a is usually reactive and its scalar value and vector angle are unknown. Even though the line were terminated for test purposes in a conductor having negligible inductance and capacitance at low frequencies, at high radio-frequencies the values might still be serious. Moreover, measurement of the impedance would be very difficult. Fortunately, however, it is not necessary to know the terminal impedance, provided the value of $Z_{max.}$ or $Z_{min.}$—pure resistances—at points near the termination can be obtained. Since we are considering transmission lines whose total length is several multiples of a wavelength, the proportionate loss over a fraction of a wavelength is negligible. Thus, referring to Fig. 25(*a*), if Z_{1A} and Z_{2A} are respectively the first minimum and maximum impedances of the line looking towards the termination, and I_{1A} and I_{2A} are the corresponding currents measured at these points

$$I_{1A}^2 Z_{1A} = I_{2A}^2 Z_{2A}$$

Moreover, $$Z_0^2 = Z_{1A}Z_{2A}$$

$$\therefore \quad Z_{1A} = Z_0 I_{2A}/I_{1A}$$

where $Z_0 = 120 \log_e (2D/d)$ for open lines
(D = distance apart of wires,
d = diameter of each wire).

Thus the output from the transmission lines is $Z_0 I_{1A} I_{2A}$. Similarly, if I_{2T} and I_{1T} are the minimum and maximum currents respectively at the transmitter end of the line, the line input is $Z_0 I_{1T} I_{2T}$. Thus the line efficiency is $I_{1A}I_{2A}/(I_{1T}I_{2T})$.

It is now desirable to obtain the attenuation of the transmission line. If I_{1T} and Z_{1T} are the associated values of current and minimum impedance looking towards the load at a point in the transmission line near the transmitter, then since the current at any point in the transmission line is given by

$$(E/2Z_T Z_0)\,(Z_0 + Z_{1A})e^{ax}e^{jBx} - (Z_{1A} - Z_0)e^{-ax}e^{-jBx}$$

where E = transmitter voltage
and Z_T = a constant,

$$\therefore \quad \frac{I_{1A}}{I_{1T}} = \frac{(Z_0 + Z_{1A})e^0 - (Z_{1A} - Z_0)e^0}{(Z_0 + Z_{1A})e^{aL} - (Z_{1A} - Z_0)e^{-aL}}$$

$$= \frac{2Z_0}{(Z_0 + Z_{1A})e^{aL} - (Z_{1A} - Z_0)e^{-aL}} \quad . \quad (5)$$

also

$$Z_{1T} = Z_0 \frac{(Z_0 + Z_{1A})e^{aL} + (Z_{1A} - Z_0)e^{-aL}}{(Z_0 + Z_{1A})e^{aL} - (Z_{1A} - Z_0)e^{-aL}}$$

Thus from (2), by substitution,

$$Z_{1T} = \frac{I_{1A}}{2I_{1T}}\left\{(Z_0 + Z_{1A})e^{aL} + (Z_{1A} - Z_0)e^{-aL}\right\} \quad (6)$$

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By substituting the value already obtained for Z_{1A}, viz. $Z_0 I_{2A}/I_{1A}$, expressions (2) and (3) may be shown in the form

$$\frac{I_{1A}}{I_{1T}} = \frac{2I_{1A}}{(I_{1A} + I_{2A})e^{aL} - (I_{2A} - I_{1A})e^{-aL}} . \quad (7)$$

$$Z_{1T} = \frac{Z_0}{2I_{1T}}\left\{(I_{1A} + I_{2A})e^{aL} + (I_{2A} - I_{1A})e^{-aL}\right\} \quad (8)$$

The efficiency of the line is then

$$\Sigma = \frac{I^2_{1A}}{I^2_{1T}} \cdot \frac{Z_{1A}}{Z_{1T}} = \frac{2I_{2A}I_{1A}}{I_{1T}\left\{(I_{1A}+I_{2A})e^{aL}+(I_{2A}-I_{1A})e^{-aL}\right\}} \quad (9)$$

or

$$\Sigma = \frac{4I_{1A}I_{2A}}{(I_{1A} + I_{2A})^2e^{2aL} - (I_{2A} - I_{1A})^2e^{-2aL}} \quad (10)$$

The value of a may be obtained from expression (7), and by using this value in (9) or (10) the line efficiency may be calculated. The efficiency shown in this form is sometimes more useful than in the form given in the earlier part of this discussion.

By applying these principles to several open transmission lines, the values shown in Table 1 were obtained.

The resistance loss may be reduced by using copper instead of silicon bronze. A further reduction may be made by using several insulated wires of small diameter instead of one larger wire for each of the two members of the transmission line. For example, a 100-lb. wire has only twice the high-frequency resistance of a 400-lb. wire of equal length. Thus, assuming that the proximity of other wires does not change the resistance values of individual wires, four 100-lb. wires would have half the resistance of one 400-lb. wire of equal length. To test the truth of this contention the 16·1-m array was fed by a group of transmission lines running along the same poles. Three insulated open lines, two of which were used normally for two other arrays, were bonded at the transmitter and the array ends. From the current values of both ends of the lines the efficiency of transmission was calculated. The result showed that the reduction in loss, namely 16 per cent, was approximately the same as the estimated reduction due to less ohmic resistance.

No method is known of estimating dielectric hysteresis losses with any degree of accuracy. To reduce insulator losses the best type of porcelain or other insulation should be used. Subject to mechanical requirements, the insulator should be as long as possible and the cross-section as small as possible to minimize leakage and reduce local capacitance at the points of attachment to transmission poles. The latter precaution is necessary to lessen small reflections due to changes in the constants of the line.

From approximate tests made by the author some years ago, the conclusion was reached that within the range of voltage gradients used for transmission-line insulators, the energy loss in insulators increases as the square of the voltage at radio frequencies. Thus a low-

TABLE 1.

Designation of line	Type of array, and wavelength (metres)	Surge impedance (Z_0) of line (calculated), ohms	Length of double-wire line, feet	Type of line	aL	$a \times 10^{-4}$	Efficiency, per cent
A	Vertical 15·59	560	500	3/18 S.W.G. copper, 9 in. apart	0·29	5·8	78·5
B	Vertical 16·38	550	1 500	400 lb. per mile copper, 9 in. apart	0·165	1·1	72·3
C	Vertical 16·1	550	2 205	7/14 S.W.G. sil. bronze, 9 in. apart	0·238	1·07	47·4
D	Vertical 24·4	550	1 695	7/14 S.W.G. sil. bronze, 9 in. apart	0·40	2·36	50·5
E	Vertical 30·1	550	2 705	600 lb. per mile copper, 9 in. apart	0·24	0·90	55·8

TABLE 2.

Designation of line, and wavelength (metres) used	Length and type of line	Resistance of 1 000-ft. single line, ohms	Ohmic energy loss expressed as percentage of input
A. 15·59	500 ft. 3/18 S.W.G. copper	55·0	7
B. 16·38	1 500 ft. copper 400 lb. per mile	27·5	16
C. 16·1	2 205 ft. 7/14 S.W.G. silicon bronze 707 lb. per mile ..	31·9	23
D. 24·4	1 695 ft. 600 lb. per mile	24·4	14·8
E. 30·1	2 705 ft. 600 lb. per mile	22·0	16·2

ANALYSIS OF LOSSES.

The losses in transmission lines may be classified under three headings: (1) Ohmic loss, (2) Dielectric hysteresis and eddy-current losses, (3) Radiation losses.

Ohmic losses are appreciable at the high frequencies used. By taking an average of the squares of the currents measured along the transmission lines, an estimate of the ohmic losses has been made. To compute values of the resistance at high frequency the methods given by the Bureau of Standards Circular No. 74 were used. The results are shown in Table 2.

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impedance line would have less loss per insulator than a high-impedance line. On the other hand, ohmic resistance losses would be increased unless the cross-section of the wires were increased.

Radiation losses from open well-balanced lines need not be great, but it will be shown presently that even small differences between the currents in the two wires of a transmission line may result in serious energy loss. Radiation losses may conveniently be considered under the following heads: (*a*) Radiation due to space separation of the two wires of the lines, and (*b*) Radiation due to inequality of current in the two lines. (*a*) and (*b*) will be dealt with together.

Consider the electric field at a point P due to a small section of a 2-wire transmission line, the point P being at such a distance as to render all lines projected from the transmission wires to P sensibly parallel [Fig. 25(*b*)].

If δL_a and δL_b are two small lengths of the horizontal transmission line L_aL_a, L_bL_b, separated by a distance D, and E is the amplitude of the electric field at the distant point P due to unit current in one of the short lengths of wire, then the amplitude of the field due to the two elements will be the vector sum of two quantities having amplitudes EI_a and EI_b but differing in phase by an angle $(2\pi D/\lambda) \sin\psi \sin(\theta + \phi)$; where θ is the co-lateral elevation of P as viewed from the transmission line, ψ is the angle between the vertical plane normal to the line and the vertical plane passing through P and the transmission line, and ϕ is the phase angle between the currents in the two wires at any two points separated by distance D. Usually $\phi = \pi$, but it is sometimes slightly greater or less than π. Using this value for ϕ, the resultant field will have an amplitude

$$E\sqrt{\{I_a^2 + I_b^2 - 2I_aI_b \cos[(2\pi D/\lambda) \sin\psi \sin\theta]\}}$$

If I_a is written $I + i$, and $I_b = I - i$, the amplitude may be shown in the more convenient form

$$E\sqrt{\{2(I^2 + i^2) - 2(I^2 - i^2) \cos[(2\pi D/\lambda) \sin\psi \sin\theta)]\}}$$

$$= E\sqrt{\left\{2I^2\left[1 - \cos\left(\frac{2\pi D}{\lambda} \sin\psi \sin\theta\right)\right] + 2i^2\left[1 + \cos\left(\frac{2\pi D}{\lambda} \sin\psi \sin\theta\right)\right]\right\}}$$

If the value of i is a small fraction of I, the power input to the section of the line considered will be sensibly the same whether the balanced or unbalanced condition is maintained.

Thus, at the point considered, the ratio of the power radiated when the currents in the two transmission wires are unequal to that when the currents are equal is

$$\frac{I^2\{1 - \cos[(2\pi D/\lambda) \sin\psi \sin\theta]\} + i^2\{1 + \cos[(2\pi D/\lambda) \sin\psi \sin\theta]\}}{I^2\{1 - \cos[(2\pi D/\lambda) \sin\psi \sin\theta]\}}$$

The minimum value of this expression is

$$1 + \frac{i^2[1 + \cos(2\pi D/\lambda)]}{I^2[1 - \cos(2\pi D/\lambda)]}$$

D is usually about 9 in. and λ is seldom less than 15 metres. Using these values the expression becomes $1 + 435\, i^2/I^2$. Thus small unbalanced currents may be much more productive of radiation than the space separation of transmission wires. In the case considered i need only be about $I/20$ in order to produce at a certain point a field strength due to the unbalanced lines twice as great as that from the balanced lines. Actually the total relative radiation will be greater than the foregoing treatment suggests. A quantitative computation could be made, but the labour involved seems unnecessary since the simple treatment already given

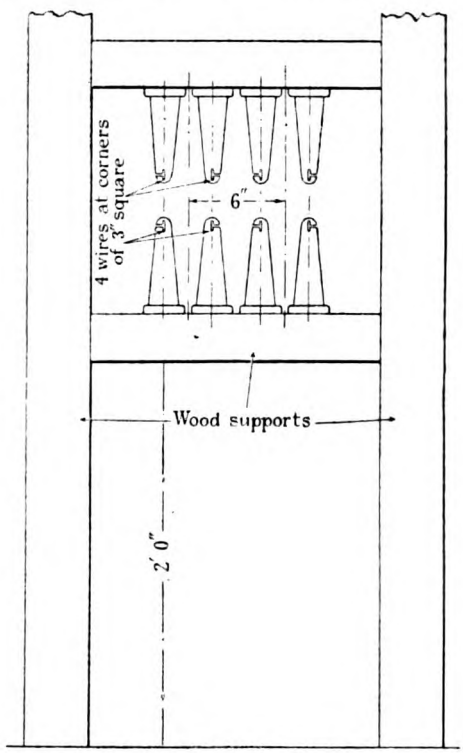

FIG. 26.—Transmission line consisting of two 4-wire leads.

achieves qualitatively the desired purpose of emphasizing the effect of unbalanced currents.

One cause of this unbalance is found in the type of array used. If, for example, the array is such that some of its members connected directly to one wire of the transmission line are nearer to the earth than the members connected to the other line, an unbalanced arrangement will result, and unless suitable counteracting precautions are taken the currents in the wires of the transmission line will be unequal. The horizontal type of array would therefore be expected to be well balanced, whilst any other type of array whose elements were not symmetrically disposed with respect to the earth and the transmission lines would be expected to be unbalanced. Both expectations have been found to be justified, and many tests have confirmed the fact that similar open transmission lines have widely different losses according to the type of array used.

To demonstrate the low radiation-losses of lines feeding horizontal arrays, Table 3 has been prepared.

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A comparison of the losses due to similar transmission lines feeding a 30·1-metre vertical array [Table 1, (E)] and a 33·25-metre horizontal array [Table 3, (G)] discloses the fact that although the former is much the shorter line, its efficiency is considerably less than that of the latter. When the measurements were made the inputs to the lines were practically the same, the type and spacing of insulation and the size of wire were identical, but the out-of-balance current was less in the more efficient lines. Nearly the whole of the loss in the lines F and G (Table 3) can be accounted for by high-frequency ohmic resistance. It thus seems apparent that the type of array employed has a pronounced influence upon the amount of energy radiated from transmission lines, and, conversely, upon the amount of "pick-up" on lines used for reception purposes.

TABLE 3.

Designation of line	Type of array	Surge impedance (Z_0) of line (calculated), ohms	Length of line, feet	Type of line	αL	$\alpha \times 10^{-4}$	Efficiency per cent
F	Horizontal dipoles $\lambda = 16{\cdot}38$ m	550	2 985	2-wire 600 lb. per mile, copper, 9 in. apart	0·175	0·59	69·2
G	Horizontal dipoles $\lambda = 33{\cdot}25$ m	550	3 460	2-wire 600 lb. per mile, copper, 9 in. apart	0·155	0·44	73·2

The transmission lines referred to in the foregoing discussion were all about 18 ft. above the ground, and as it was thought possible that radiation due to unbalanced currents might be reduced by building lines near to the earth, comparative tests were made upon similar 2-wire 200-lb.-per-mile transmission lines built 18 ft. and 2 ft. above ground. The results showed that the efficiency of the lower line was not greater than that of the higher, due probably to ground losses neutralizing any reduction in radiation losses.

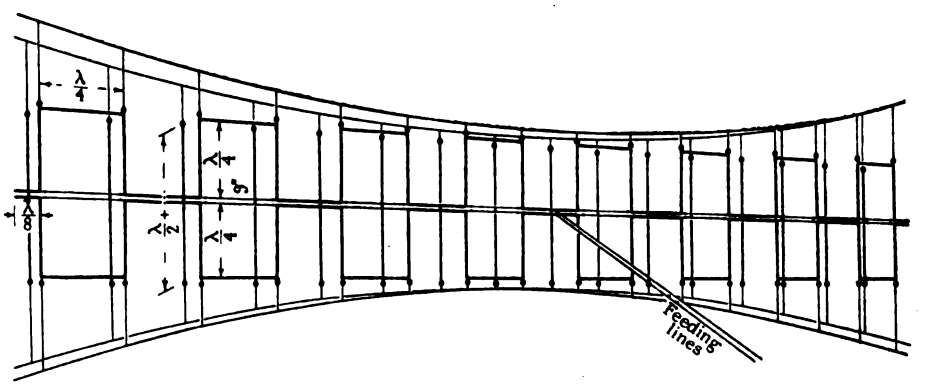

FIG. 27.—Double key vertical array with reflector.

MULTI-WIRE TRANSMISSION LINES.

It has already been pointed out that a given section of conductor can be most effectively used, as far as ohmic losses are concerned, by being divided into several insulated paths. A multi-wire line 1 900 ft. long was built to take advantage of this fact. It consists of four 200-lb.-per-mile copper wires, built 2½ ft. from the ground (Fig. 26), the wires of each limb being insulated from each other but joined together at their extremities. When connected to a type of array (Fig. 27) that gave rise to unbalanced currents, the efficiency of the line was 71 per cent on a wavelength of 20·8 metres. The calculated ohmic loss is 16 per cent, leaving 13 per cent for radiation and other losses. The ohmic resistance could, of course, be reduced to a low value by increasing the number of wires. A line 1 100 ft. long consisting of a pair of 200-lb.-per-mile wires fixed 18 ft. high and spaced 9 in. apart was connected to an exactly similar array, except that the working wavelength was 16 metres approximately. The loss in this case, excluding the resistance loss, was 34·5 per cent, which compares unfavourably with the loss in the multi-wire line.

The measured surge impedance of this multi-wire line is 190 ohms, being slightly higher than that of an equivalent line consisting of parallel tubes of 3 in. diameter spaced 6 in. between the central axes. As the comparative cost of a concentric tube line and a multi-wire line is 5 : 1 approximately, a multi-wire line forms an economical substitute for a tube line. It has the additional advantage of simplifying the connections to arrays. For example, pairs of wires of a 2-4-wire line can be selected to connect to each of the units of a 4-unit array. The spacing of the grouped wires can readily be adjusted so that at the position where the various pairs branch off to the array, correct impedance values are obtained. Such an arrangement has advantages over the usual divided feed, inasmuch as it eliminates several

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junctions and thus reduces the number of impedance matching connections. The necessity for matching impedances correctly arises more on account of the reduction in power delivered to the aerials than from the resulting losses in lines, if this precaution is not taken.

To summarize, the conclusion has been reached that losses in open transmission lines are caused by:—

(1) Unbalanced currents in the line wires.
(2) Ohmic high-frequency resistance.
(3) Space separation of the wires.
(4) Dielectric losses in insulators.
(5) Reflections from points where the line constants change.

Methods have been suggested whereby the losses can be reduced and it has been shown that, contrary to the usually accepted belief, open transmission lines—particularly those used in conjunction with symmetrical types of aerial systems—may have low electrical losses.

The thanks of the author are due to the Engineer-in-Chief of the British Post Office for permission to publish this paper; to Lieut.-Colonel A. S. Angwin, D.S.O., M.C., without whose generous encouragement this work would not have matured; and to his colleagues—particularly Messrs. F. C. Carter, A. H. Mumford, T. Davies and P. L. Barker—for their enthusiastic assistance.

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Discussion before the Wireless Section, 3rd December, 1930.

Mr. G. Shearing: The paper is an important contribution to our knowledge of short-wave aerial arrays. Especially valuable is the manner in which the author has dealt with the practical design and economics of the aerial systems. I note that the results obtained by the Post Office Engineering Department confirm the conclusions of other workers, notably Franklin and T. L. Eckersley in this country, as regards the value of low-angle radiation. The Naval Signal School, Portsmouth, some time ago carried out a series of experiments using small arrays on transmissions from Horsea at varying angles to the horizontal, and although in a few isolated cases good received signal strengths were recorded for higher-angle radiation, the average of all the reception results reported from the Far East was greatly in favour of lower-angle radiation, making an angle with the horizontal certainly not greater than 15° for wavelengths between 15 and 40 m. On pages 17 and 19 the author infers that the earth is a perfect conductor, whereas many experimenters have assumed that the earth has considerable resistivity at these high frequencies. For the aerials described, with the lower ends at a height of the order of one-quarter of a wavelength from the ground, I should have expected the earth loss to be appreciable, and I have usually assumed that this height should be of the order of one wavelength or more above the ground for reasonable reduction of earth losses. I should like to ask whether the author has carried out any actual radiation-resistance measurements on the aerials referred to in the curves of Fig. 12, for comparison with the calculated values plotted in the figure. The reference on page 20 to lateral deflection for the case of a narrow beam refers, I assume, to the displacement of a given beam from the great circle bearing normal to the array. It is conceivable, especially for transmissions east and west where change from a light to a dark zone may occur, that lateral dispersion of the beam may be produced, and I should

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like to know whether any evidence of this has been obtained for the transatlantic or other short-wave services, whereby the service area for the main beam is greater than the theoretical area as calculated from the angle of the beam at the actual array. The portion of the paper dealing with the economical size of aerial array is interesting, but I suggest that the author should expand the second paragraph of this section so as to deal more fully with the comparison of aerial arrays. As regards the relative effects of reflector curtains at three-quarters and one-quarter of a wavelength from the exciter curtains, I am not clear whether the reflector curtain has the same total width as the exciter wires. Perhaps the author will state the spacing between the reflector wires. In connection with the improvement obtained by increase of the number of reflector wires, is the explanation that as the wires come closer together the phase of the induced currents becomes more uniform throughout the curtain? Also, does the Post Office Engineering Department contemplate using reflector curtains with three-quarter wavelength spacing from the exciter wires? With regard to the interesting polar diagrams shown in Figs. 23 and 24, has the horizontal type " T.W." array with a calculated direction of radiation inclined at 10° to the horizontal proved to be more effective than the vertical type with a direction at 0° to the horizontal? The fact that the open type of transmission line can be made efficient by arranging for a perfectly symmetrical design throughout the aerial system is important, in view of the considerable saving which results in the capital cost of the aerial feeder system.

Dr. R. L. Smith-Rose: I should like to join Mr. Shearing in expressing my appreciation of this paper, on the subject of which very little has been published in this country. One is apt to obtain the impression that all the work in connection with antenna arrays is being done abroad, since most of the published work has emanated from Germany, Japan and America. The paper brings to light the second of two important advantages associated with the use of short waves in radio communication. The first advantage is that the radiation efficiency of a transmitting aerial is much higher for short than for long waves, and the second is that one can effectively increase the useful radiation from the transmitting aerial by employing an antenna array. From the figures given in the paper it appears that a power ratio of the order of 100 to 250 times could be obtained with a suitably designed antenna array, i.e. a 10-kW beam station using a short wavelength becomes equivalent to a 1 000- or 2 000-kW station distributing its radiation in all directions. A considerable amount of theoretical work underlying the design of antenna arrays has been done in this country at the National Physical Laboratory under the auspices of the Radio Research Board and I would suggest that reference to some of this—particularly to a group of three papers recently published in the *Journal**—be inserted in the paper. In the first of these, by Messrs. R. M. Wilmotte and J. S. McPetrie, the substance of Fig. 15 of this paper was contained, and it was definitely pointed out that the optimum spacing between an excited antenna wire and the reflector wire for maximum forward radiation was either 0·33 or 0·85 of a wavelength instead of the previously accepted values of one-quarter or three-quarters of a wavelength. This was a theoretical deduction, and I am interested to note that it has been borne out by some German theoretical work and experimentally confirmed by two American investigators. I would also point out that additional experimental confirmation was given by Messrs. L. S. Palmer and L. L. K. Honeyball* in a paper published in the same month as the paper by Englund and Crawford appeared in America. In view of the fact that the optimum distance has been found to be one-third of a wavelength, I should like to ask whether the author has carried out any experimental work with this spacing. In the case of the arrays given in the paper the spacing is definitely referred to as one-quarter of a wavelength, and I am at a loss to understand what advantage, if any, is gained by this arrangement. In view of the confirmation of the fact that the optimum distance is one-third of a wavelength, it will be interesting to see whether arrays will in future be built with this spacing instead of one-quarter of a wavelength. I share Mr. Shearing's views on the effect of the conductivity of the earth. During the past few years our knowledge of the conductivity of the earth and of its effect on radio-wave propagation has considerably increased and it is possible to take the earth into account in making calculations on antenna systems. Incidentally, Wilmotte in a very recent paper† stated that he had been able to allow for the effect of the conductivity of the earth as measured on the site of the transmitting station, and thus to correct the calculated polar diagrams and obtain better agreement with his experimental results. With regard to the question of the best angle of elevation at which to project the beam, it at first seems obvious that in order to send a beam over a distance of several thousand miles it is necessary to project it at an angle of 10° or 20°, or more, to the horizontal. On further consideration, however, the reason for expecting the optimum direction of radiation to be approximately horizontal becomes apparent. For example, Fig. A shows a section drawn to scale of the earth with its radius of 4 000 miles, and the ionized layer at a distance of 100 miles above it. A wave starting from the transmitting station at a large angle to the horizontal will come down after travelling 200 or 300 miles. On the other hand, a wave which leaves in a horizontal direction passes the earth again at grazing incidence, and, after a second reflection, comes down at a receiving station approximately 3 500 miles away. This is about the distance between London and New York. Thus the wave which stands the most chance of getting from London to New York is one which leaves in a direction approximating to the horizontal. Any other wave would be reflected many more times on its passage, and would suffer more loss by the repeated reflections at both the earth and the ionized layer. As the earth is an imperfect conductor it is not possible to project a beam exactly horizontally, and it is therefore necessary to have an elevation of a few

* *Journal I.E.E.*, 1928, vol. 66, pp. 949–967; also *Proceedings of the Wireless Section*, 1928, vol. 3, pp. 182–200.

* *Journal I.E.E.*, 1929, vol. 67, p. 1045; also *Proceedings of the Wireless Section*, 1929, vol. 4, p. 152.

† *Journal I.E.E.*, 1930, vol. 68, p. 1191; also *Proceedings of the Wireless Section*, 1930, vol. 5, p. 235.

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degrees, just sufficient to free the radiation from the effects of the finite conductivity of the earth. I do not quite understand the author's use of the term "radiation resistance" in his reference to the work of Pistolkors. I gather that Pistolkors defines radiation resistance as the radiated power divided by the square of the current; but in view of this, and taking into account Fig. 12, I am at a loss to understand why it was considered an advantage to choose the point at which the radiation resistance was a minimum. Consideration of the definition leads me to think that the condition of maximum radiation resistance is what is required. I use the term here in its broadest sense, taking it to include the effect of neighbouring wires and of the earth. Perhaps the author could enlighten me on this point. From Fig. 7 it would appear that the quarter-wave aerial gives the same radiation in a horizontal plane as a half-wave aerial, but this is surely only true if the current in the former case is twice that in the latter. I suggest that Fig. 24 be rotated through 90° so as to be presented in the same way as Fig. 23.* I gather that one of the wavelength, has contained within it a horizontal array for a 30-m wavelength; the other, a 33·5-m vertical array, has suspended within it on the same masts and cross-arms a horizontal array with a 25-m wavelength. The actual loss measured on the second of these combined arrays is only of the order of 0·5 decibel, a negligible loss compared with the saving in cost. The author deals with the question of height of arrays as affecting signal strength at a distance.. The fortunate circumstance that two 820-ft. masts were available at the Rugby radio station enabled a large-scale experiment to be carried out which has some bearing on the point mentioned in the paper. Utilizing these masts to carry a triatic, a combination of vertical and horizontal radiating elements was erected which consisted of 64 half-wave horizontal and 128 half-wave vertical elements, arranged in one large network about 1 020 ft. long, 400 ft. in depth and raised 50 ft. above the ground. These aerials could be fed simultaneously, so that they radiated from both the horizontal and vertical elements, or alternatively the horizontal and vertical

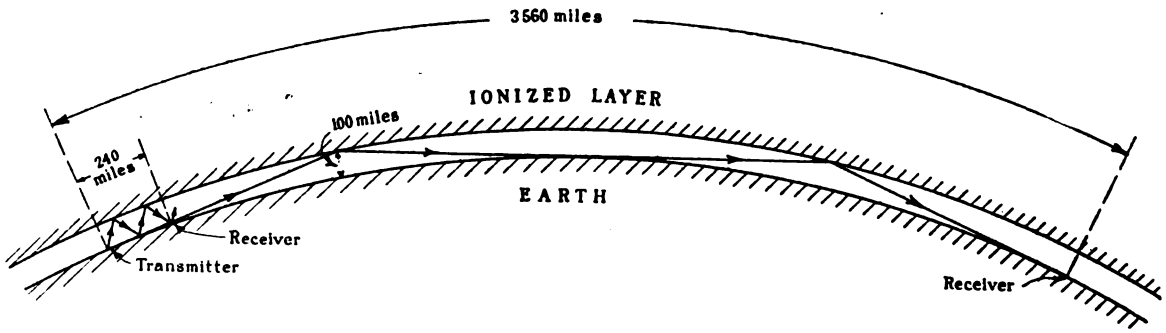

FIG. A.—Section of Earth, radius 4 000 miles, and ionized layer at a height of 100 miles, showing the paths of waves between a transmitting station, and two receivers at distances of 240 and 3 560 miles.

great advantages of the type of array which the author has developed is that the use of high and expensive masts is dispensed with, and one is able to effect a great economy in the general construction of the array. Incidentally, on close study of the paper one appreciates that the author has done great service in designing an economical and efficient type of self-supporting mast. What are the relative advantages of the horizontal as compared with the vertical type of array? I am unable to draw a definite conclusion from the paper on this point.

Col. A. S. Angwin: It is refreshing to see a mathematical treatment of this subject applied so directly to economic considerations. The author stresses the economic aspect of the large area of site occupied by array systems, particularly where concentration of services and alternative wavelengths have to be considered. In the past the tendency in laying down transmitting and receiving aerial systems has been to prolong them in long lines in order to prevent the possibility of mutual interference. The tendency now is in the direction of erecting more than one array on the same supporting structure, and it is interesting to note that the Post Office Engineering Department have now two groups of arrays in which this device has been successfully adopted. One, a vertical array for 42·92-m

* This has been done for the *Journal*.

elements could be fed independently. The wavelength was 27·8 m, and a comparison was made with an aerial using the same wavelength and the same number of horizontal elements, but erected on 180-ft. masts. It was interesting to observe that, in spite of its much greater height, the actual results received in Australia from the first aerial with the horizontal elements alone excited were slightly worse than those obtained from the low aerial system, and the efficiency of the horizontal and vertical aerials together on the high masts was definitely worse than that of the horizontal system on the lower masts. This emphasizes the point to which the author refers, that beyond a certain limit the intensive narrowing of the beam may tend to be prejudicial to, rather than to improve, the effect received at a distance. I am not clear with regard to the assumption made by the author that the earth is a perfect conductor. Measurements have been made of the conductivity of the earth at the frequencies dealt with in this type of array, and figures varying from 6° to 10° are given for the wave tilt. One might assume from this that in the case of horizontal-type arrays there would be some measurable field on the ground, but the author's experiments indicate that this is not so. I think that the apparent contradiction between these two facts should be the subject of further research, or at any rate of

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mathematical analysis. More information as to the apparent values of the earth's conductivity at high frequencies would be a valuable contribution to the subject.

Captain J. W. S. Dorling: I should like to ask two questions, the first of which is connected with the economics of the subject. The author points out that a gain in signal strength may be obtained by increasing the height of the masts and using a more extensive array. Has he considered the economic aspect of getting the same increase of signal strength by an increase of power at the transmitter? Secondly, has he considered the question of widening the beam and thus communicating with a zone rather than with a definite point on the earth's surface? The case I have in mind is the ship-to-shore telephony services in operation on the North Atlantic route. It seems to me that in some instances a wide-angle beam would be an advantage for getting signals to the ships, and possibly for receiving signals from those ships. It might perhaps result in a decreased cost of the apparatus to be installed in the ship, or, what is even more important in some cases, a less complex apparatus.

Mr. F. S. Barton: We at Farnborough were associated with the work done there by Mr. Wilmotte to which reference is made in the paper, and it would be interesting to us to know how valuable this was and whether more work of a similar type will be necessary. The author regards arrays almost entirely from the point of view of how strong a signal he can obtain for a given amount of money, whereas in the Royal Air Force we want mobility also. The earlier types of arrays and transmission lines do not seem to be adaptable to mobile use. The introduction of the open-wire transmission lines and quarter-wave matching lines described by the author, however, seems to make possible the use of an array in a mobile station. Inspired by the work of the Post Office Engineering Department, we have made a brief study of the subject and have erected a simple Koomans array consisting of four half-wave dipoles, built up almost entirely from mobile-station equipment. This attempt was made primarily in order to find out whether one could realize in practice, with this simple mobile equipment, a calculated design, and secondarily to ascertain whether this design was of any use for services of a mobile nature.

Mr. A. J. Gill: I should be glad of further information regarding the relative advantages of the various arrays. A peculiar feature of the Sterba array is that it is a closed circuit and can thus be energized by low-frequency currents in order to prevent the formation of sleet under certain weather conditions. With regard to the Koomans array, especially the horizontal pattern, has the author considered the possible advantage of placing the vertical feeders a little nearer together, and allowing the ends of the aerial to overlap? This would increase the current concentration and enable more stacks of aerials to be placed in a given span between masts. I think that there is a possibility of increasing the efficiency of the mast arrangement by this method. In some of the calculations for Fig. 13 the author has taken the value of the dead-loss resistance of a half-wave element to be about 25 ohms. I should like to ask what reasons he has for assuming this rather high value and what effect any variation of the figure will have upon the results. With regard to the multi-wire transmission line shown in Fig. 26, I suggest that a considerable advantage would be gained by transposing the lines. The impedances of the individual wires apparently amount to about 500 ohms in the case of inner wires, and 620 ohms in the case of outer wires. This should lead to uneven distribution of current in the four wires comprised in each of the lines. I should like to know what accuracy can be assumed in the figures given for the efficiency of lines. In our early work with transmission lines, having, as we thought, balanced the line, we attempted to take some measurements; it was found, however, that apparently far more power was being received at one end of the line than was being fed in at the other, and we had to abandon the attempt. The results given in the paper are apparently based on certain calculations of the value of Z, and I should like to know the author's estimate of the accuracy of these.

Squadron Leader D. H. de Burgh: The point at issue is the efficiency as opposed to the directional properties of the array. At Farnborough I am using an array consisting of three ordinary mobile masts with guys, ordinary telegraph insulators (which the author regards as not good for this purpose), and 200-lb. wire; the preliminary results with this arrangement appear to show that the wavelength of the array is that for which it was designed and that the distance apart of the matching lines agrees with the calculated value. There is a discrepancy, however, in that a current node exists about 3 ft. from the correct position at the end of the matching lines. Although this does not appear to make much difference to the current in the array itself, I naturally cannot get exactly equal currents in the line. The discrepancy could probably be corrected by adjusting the length of the half-wave lines joining the matching lines to the two halves of the array. In view of the fact that there may be some doubt as to the calculated value of the impedance of radio transmission lines, what is the present position regarding the possibility of measuring the impedances of such lines at high frequencies?

Mr. F. C. Carter (*communicated*): There are one or two points in the paper on which I should like to comment. In the first place, when comparing various methods of grouping the radiating elements in the several array systems the author assumes that the same current is present in the centre of each element. This may be true for certain types of arrays, such as the horizontal Koomans shown in Fig. 3, where the pairs of horizontal conductors constitute equal parallel resistive loads at the bottom of the vertical feeders, and therefore (neglecting the small ohmic loss in these feeders) all carry equal currents. In other types of arrays, viz. the Sterba (Fig. 4) and Marconi-Franklin (Fig. 2), certain of the antennæ are in series with one another, so that those nearer the energizing source carry larger currents than those more remote. It is obviously difficult to determine this distribution in the case of a vertical-type array, but from some tests made on an 8-section double Bruce array the current in the radiators near the energizing feeder was found to be approximately three times that in the outermost radiator. If such an unequal current distribution occurs

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in a vertical-type array, there is a substantial reason for limiting the economic height of beam arrays, since the higher the radiators the less effective they become. The author, in common with many other writers on this subject, tacitly assumes that the best vertical spacing of the tiers of an array is one half-wavelength. This is probably because with such a spacing it is easy to arrange the bends in the wires so that all the radiators are correctly phased. As an alternative to this half-wavelength spacing, I suggest that a vertical spacing of one-quarter wavelength is worthy of consideration with a horizontal type radiator. This could readily be accomplished by using one-half of a double Bruce type array (shown in Fig. 27) turned through 90° and energized at the bottom. A number of such panels side by side would comprise the array. By suitably adjusting the length of the uppermost vertical element another closed-loop type array is obtained. By these means twice as many radiators can be arranged while still using the same height of tower. The spacing of the reflector curtain behind the array has long been a controversial subject, and perhaps the author dismisses it rather lightly. Consider a simple array and reflector; the field in front of the array is the sum of two vectors, one proportional to the current in the directly energized radiators, the other proportional to the current in the reflectors and displaced from the first vector by an angle depending on (*a*) the distance between the curtains and (*b*) the phase relationship of the currents in the aerial and reflector. At distances of the order of one quarter-wavelength the current in a radiation-excited reflector increases rapidly with a small decrease in spacing, with only a small resultant change in phase of the current, provided that the reflector* is correctly tuned. It is thus possible for the resultant vector sum to be greater than that obtained when the spacing is one quarter-wavelength. I have carried out some experiments which, although unfinished, showed the optimum spacing for maximum forward radiation to be 0·21 wave-length. It is of interest to note that Tatarinoff* advocates a spacing of 0·2 wavelength for the rear reflector of a parabolic reflector. The major portion of the paper deals with transmitting arrays, and receiving arrays are different in many respects. Among other points it would appear that there is an optimum width of receptive angle in the vertical plane, determined, of course, by the number of tiers. Perhaps the author would indicate whether he thinks it preferable to make this angle nearly zero, or alternatively, to use a beam having a wide receptive aperture in the vertical plane—as advocated by Schelleng.† A more convincing method of drawing Fig. 13 would be, I think, to show the effect of a similar number of radiating elements distributed either vertically, horizontally or in planes.

Mr. H. Chireix (*communicated*) : The aerial referred to by the author as the " Sterba " type is covered by my own patent. I cannot agree that a line consisting of coaxial conductors can cost five times as much as one having two parallel conductors. Moreover, I myself had already noticed the critical length of the wires which constitute the component parts of reflectors.

* *Zeitschrift für Hochfrequenztechnik*, 1926, vol. 28, p. 117.
† *Proceedings of the Institute of Radio Engineers*, 1930, vol. 18, p. 913.

Captain P. P. Eckersley (*communicated*): I will ask the author some purely economic (as opposed to purely technical) questions. I am encouraged to do so since he has expressed the results of his equations in pounds and shillings. The economic answer would seem to be most logically expressed in terms of the cost of establishing a field of sufficient strength at the point of reception. The design of aerial is surely only one of many factors in this calculation. I suggest that the cost of the site for the station, the cost of power, the cost of maintaining the land lines, and the cost of maintaining and building the aerial structure, should all be taken into consideration, and the proportion which the cost of each bears to the total expenditure should determine its importance in the total economy considerations. Obviously, also, the field at a distance is proportional both to the radiation efficiency of the aerial (useful power divided by the power in the aerial) and to the power input to the transmitter. With a given aerial and given conditions, the field at a distance is increased directly as the power is increased. Would it be better to build a less efficient aerial and employ more power, or vice versa? (It is obviously the beam aerial that is here referred to, and the term " efficiency of aerial " refers to the efficiency of a beam aerial.) Thus, while the rather complicated mechanical design of aerial described may make the device, as such, so much more efficient, economically and technically, in creating a field at a distance, might not transmitting stations, disposing of simpler aerials, requiring, however, greater power, be still more efficient when efficiency is defined as the cost of establishing a field at the point of reception? Further, if this more general definition of economy be adopted, is there so much to be gained by using the design of aerial described, seeing that it is probably more difficult to construct and maintain, although more economical to support? The answers to these questions would seem to indicate the real relative efficiency of transmitting stations considered as complete units. The aerial is, after all, a component of a station and perhaps a not relatively expensive component. It may merit less consideration than other components and may be made sufficiently efficient by using greater power. A contributor to the discussion asked why the radiation resistance increased as the end of the aerial was brought nearer to the ground, and why, if that were so, it was not desirable to bring the end near the ground. I hope the author will emphasize his answer to that question. It is remarkable that this point, fundamental to an appreciation of even the essence of this and other papers on aerial design, is so frequently missed or not understood. A clear explanation was first given by an American, Stuart Ballantine.

Mr. C. S. Franklin (*communicated*): Before I designed the aerials for the imperial beam stations I investigated theoretically their possibilities. Assuming only Huygens's principle of propagation from a plane wave-front of definite area, Neumann's formula for the field at a distance from an element of current, and that the current in the aerial could be treated as a uniform sheet, I plotted the field intensity for a spherical surface of large radius compared with the aerial. Integrating the mean square of this field for the whole surface gave

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a figure proportional to the energy passing through it. By repeating this procedure for various types of aerial I obtained comparative figures for several forms of aerial of the total energy which must be radiated to produce equal fields in a particular direction. I chose for comparison an aerial in which the radiation is confined to a single wire of small length in comparison with the wavelength. The result of the calculations showed that for aerials of this type having an aperture in both directions of upwards of one wavelength, the energy magnification figure which could be obtained per square of wavelength was 9·6. Compared with a half-wave aerial this magnification figure was approximately 8·7. Calculations were made to find the effect of the current being concentrated in the several wires instead of treating an aerial as a uniform current sheet. These showed that no material change resulted, provided that the wires were half a wavelength or less apart, and, as regards the vertical distribution, that the points of maximum current in phase were not substantially more than half a wavelength apart. These figures were published in a paper* by Marchesi Marconi in 1924. In this investigation I did not take the earth effect into account, as it was at that time, and still is, a very controversial subject. If the effect is taken into account it will modify the polar curve of any particular aerial in the vertical plane to a degree depending on the constants chosen, but I do not think that it will alter materially the comparative figures between various aerials. Using the methods described, I find that the comparative gains of aerials B and D (Fig. 13) when used with reflectors are 14 and 17·4 decibels respectively. I have not completely investigated aerial C, of the same figure, but from an inspection of the modification of the polar curve due to the introduction of the second line of radiators half a wavelength in front and 180° out of phase I find the gain of C over B to be quite small. As aerial C is the one specially recommended by the author, I think it important to establish whether his figures are correct or not. At present I disagree with them. There is a type of beam array consisting of a row of radiating aerials the current in which is so phased that the propagation is along the line of the aerials. Such a system gives a radiation confined within a cone, and the angle depends upon the length of the line used. When each element is a single wire a considerable magnification or gain can theoretically be obtained. The author's "T.W." aerial with a reflector is partly of this type. When, however, the units in such an aerial instead of being single wires are multiples of wires having a considerable extension both horizontally and vertically, with respect to the wavelength, the gain obtained by these arrangements in line one behind the other soon becomes very small and for one wavelength extension the gain of 3 units over 2 is hardly worth while. If I am right, the author's economic argument fails. This argument also appears to be unsound from another point of view, because the relative cost of aerials C and D, which are of two different types, are compared. I think that it would be more useful to give the relative gains and costs of C and D with D converted to the same type as C but keeping the same height as before. The author, in his description of the "T.W." vertical type of aerial, mentions the radiation from the horizontal members. I believe I originated the scheme of folding a long wire so as substantially to suppress radiation from alternative half-wavelengths, and so obtain concentrated radiation in a direction at right angles to such a wire. This principle made it possible to extend the beam aerials vertically, and to obtain in a simple manner concentration in the vertical plane. A number of different aerials utilizing this principle have been designed, and the method adopted for the imperial beam stations was only one of several possible arrangements. The principal reasons for its adoption were:—(1) Practically all horizontal wires, and their attendant disturbing effects, were eliminated. (2) The design enables the feeder distribution system and attachment points to be standardized for a wide band of wavelengths. (3) Damage to one or two elements of the aerial does not shut the station down. Since the original beam-station aerials were designed developments have taken place which, while maintaining the above characteristics, give a more effective use of the area of the aerial and a much improved polar curve in the vertical plane. The "T.W." array described by the author partly makes use of this principle, but I do not think that any considerable suppression of radiation from the horizontal members is obtained. I agree that side radiation from these members will be small, but there will be considerable radiation at angles some 50° above the horizontal in both forward and backward directions, and this will cause the loss of a considerable amount of energy. These loops considerably modify the distribution shown in Fig. 23. The author's remarks on transmission lines are interesting, and I agree with him that these are as important as the aerial. I was responsible for the adoption of the concentric-tube feeder, and, although the cost was high, I think that the decision was justified. Before deciding upon the concentric-tube type of feeder for the beam stations, I made experiments with the open twin type of feeder and also the twin feeder in an earthed case. The open-type feeder was not adopted, principally because of its inherent instability, although reasonable efficiency could be obtained with it. The concentric-tube feeder, on the other hand, proved both efficient and stable. The concentric tube need not differ greatly from the twin open type as regards cost, as the feeder can be made in the form of wire cages. Practical considerations in many parts of the world, however, decided that the feeder should take the form of actual tubes. Mr. E. Green, who helped a great deal in the development of the aerials and feeders, worked out theoretically the properties of the latter and showed that the maximum and minimum values of impedance, voltage and current occur at quarter-wave intervals along the feeder; he also found that the impedance at points of maxima and minima was purely resistive. These facts were made use of in the feeder distribution system of the beam stations, and they enabled the impedance matching devices to be eliminated at certain junctions.

Mr. E. Green (*communicated*): As I assisted Mr. Franklin during the development of the Marconi type of beam aerial and feeders, I should like to make a few comments on the following items in this paper:—

* *Journal of the Royal Society of Arts*, 1924, vol. 72, p. 607.

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Reflectors.—The normal front-to-back ratio of field strength of an ordinary exciter curtain and properly adjusted reflector curtain should be 15 : 1 or more. The author's figure of 12 : 1 mentioned when the paper was read compares reasonably with this. Data obtained from the imperial beam stations are not very relevant as these reflector curtains were erected to calculated dimensions, when the data on reflector wires were more meagre than they are to-day, and it was not possible to do much experimental work on stations being erected to contract.

Transmission Lines.—Tables 1, 2 and 3, giving the efficiencies of open-wire lines feeding vertical and horizontal aerials, are interesting. They show the risk of having large unknown radiation losses on open-wire lines. The ordinary erecting engineer can hardly be expected to detect and correct such losses. I find the author's remarks and experimental results on multi-wire feeders rather puzzling. He states that he connected three similar, independent transmission lines in parallel, and found (as he expected) a gain in efficiency due to the lower ohmic resistance of the line. Assuming that in each case the line is correctly terminated, there should, however, be no difference in efficiency between a single 2-wire line and any number of similar lines connected in parallel. For example, if three lines are put in parallel the resultant line resistance (which may include both ohmic losses and insulator losses) is reduced to one-third of the value for a single line. But the surge impedance, and therefore the correct terminal load resistance, is also reduced to one-third of its value for a single line. Hence the ratio of load resistance to line resistance is unchanged, and therefore the efficiency should be unchanged. The author's example of a multi-wire line is not quite the same as a number of independent lines in parallel, on account of mutual reactions between the wires, but the figures given indicate that the same amount of copper put into a single-wire line would result in a higher efficiency. The surge impedance of the multi-wire line is given as 190 ohms. The line resistance of the multi-wire line cannot be less than $\frac{1}{4}R$, where R is the effective resistance of a single-wire line of the same gauge. If instead of the multi-wire line we put the same amount of copper into a single-wire line, its resistance would be $\frac{1}{2}R$, and its surge impedance at 9-in. spacing 550 ohms. Therefore:—

$$\frac{\text{Ratio of load resistance to line resistance of heavy-gauge line}}{\text{Ratio of load resistance to line resistance of multi-wire line}} = \frac{550}{\frac{1}{2}R} \cdot \frac{\frac{1}{4}R}{190} = 1{\cdot}45$$

This means that the single-wire line of heavy gauge could be at least 1·45 times as long as the multi-wire line and yet give the same efficiency.

Impedance Matching Devices.—Patents Nos. 274970 and 282905 taken out by Mr. Franklin and myself in 1926 give various ways of matching impedance at the junction or termination of transmission lines. These include combinations of coils and condensers as well as the devices favoured by the author, i.e. the use of a portion of transmission line of appropriate surge impedance as a transformer. The elementary theory of transmission lines at radio frequencies which led up to these devices was published in 1928.*

* *Experimental Wireless*, 1928, vol. 5, p. 304.

Mr. T. Walmsley (*in reply*): Several speakers have commented upon my assumption of a perfect earth for the purpose of calculating gains of arrays, and the suggestion is made that such an assumption is not justifiable. The question must therefore be asked: What is the alternative? Certainly not a perfectly insulating earth. From purely electrical considerations the dielectric constant and conductivity of the earth should be taken into account. Several disturbing factors, however, militate against the practical application of theoretical principles. Firstly it is questionable whether the dielectric constant of the earth for short waves is known with even approximate certainty; secondly, experimental investigators are not agreed that the value of the dielectric constant remains constant with change of wavelength*; thirdly, the moisture content of the earth affects both its conductivity and dielectric properties. Even assuming an exact knowledge of the constants of the earth, the practical difficulties encountered in calculating the polar diagrams for the purpose of comparing the gains due to various groupings of wires are very great. To instance but one difficulty—for each particular angle of elevation considered the amplitude and phase angle of the reflected ray changes. Unfortunately, nebulous theories and vague assumptions do not assist the practical designer. He is compelled to accept the most workable hypothesis and if that hypothesis yields results that accord reasonably well with observed facts, he is justified in his assumption as a practical expedient. Such is the state of affairs in the present case.

In reply to Mr. Shearing's queries I have not verified practically the curves of Fig. 12. The displacement of a given beam refers to deviation from the great circle bearing normal to the array; the few measurements made in America on the transatlantic Rugby beam show that the calculated width of the main beam agrees reasonably well with the actual width, but as far as I am aware there have been no extensive tests to prove this matter. With regard to Mr. Shearing's suggestion that fuller comparison of arrays should be made, he will probably find that by utilizing the curves of Figs. 13 and 14 and applying the other information given on page 20, he will be able to compare several types of array. In all arrays used by the British Post Office the reflector curtains have approximately the same total span as the exciting curtain. Usually the spacing between adjacent reflector wires is half a wavelength, but in some cases when two exciter curtains are used in conjunction with one reflector curtain the reflector wires are spaced one-quarter of a wavelength apart. Such an arrangement gives a slightly better reflector action, due probably to an increase in the product of current in each reflector wire and the number of the reflector wires. It is not anticipated that the Post Office Engineering Department will make a general practice of using a separation of three-quarters of a wavelength between exciting and reflecting curtain when only one exciting curtain is used.

* J. P. Ratcliffe and F. W. G. White: *Philosophical Magazine*, 1930 vol. 10, p. 667; also N. J. O. Strutt: *Elektrische Nachrichten Technik*, 1930 vol. 7, p. 387.

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With regard to the relative effectiveness of horizontal and vertical arrays, the evidence shows that, for a given capital expenditure, the horizontal type is the better.

Dr. Smith-Rose points out that credit should be given to Messrs. R. Wilmotte and J. S. McPetrie for their theoretical work, and I regret that inadvertently I omitted reference to their published papers which take priority to some of the papers of other investigators. With regard to the question of the best distance between a single reflector and a single exciter wire, some tests have been made by Mr. F. C. Carter of the Post Office Engineering Department. These are referred to later in the discussion. I have carried out some experimental investigations on the effect of making the spacing of reflector and exciting curtains a little greater or less than a quarter of a wavelength, but these tests were not exhaustive since, when it was ascertained that a 12 : 1 ratio of front to back field strength was obtainable with a quarter of a wavelength spacing, it was considered of little economic importance to proceed further with the matter. The problem of spacing of curtains is obviously different from that of spacing of two wires. Dr. Smith-Rose's remarks concerning Wilmotte's ability to correct the calculated polar diagrams seem to me to display an optimism not shared by Mr. Wilmotte himself. The views of Dr. Smith-Rose on the best angle of projection of a beam are interesting but hardly accord with fact. He appears to suggest that 3 500 miles is the distance of optimum signal strength, but the reports from ships steaming between England and America do not substantiate that belief; signal strength appears to be about equal over large areas. The reason why the minimum value of radiation resistance is taken is explained at the bottom of the second column on page 19. As the aerial is raised, the volume of the solid polar diagram decreases for a constant current value in the aerial and a constant field strength at a distant point in a horizontal direction. Thus the power input and hence the total radiation resistance decrease, whilst the distant field strength remains constant. I support Dr. Smith-Rose's suggestion to rotate Fig. 24 through 90°, and I have arranged to do this for the *Journal*. The query concerning the relative advantages of the horizontal and vertical types of array was answered in my reply to Mr. Shearing.

Colonel Angwin mentions the tests made at Rugby to compare two arrays similar in type, but one having eight lines of horizontal radiators, one above the other, and the second having four lines of horizontal radiators. In both cases the spacing between lines was half a wavelength, but the higher aerial had half the span of the lower. The same transmitter was used for the tests and, by means of a throw-over switch, the aerials were alternately energized at intervals of a minute or two. When all connections had been made, the lower array was shown to be definitely superior to the higher. The explanation probably lies in the fact that a greater gain is theoretically possible by increasing the span rather than the height, and is also due to the difficulty in feeding current into the upper radiators of an array.

The question, raised by Captain Dorling, of the economic advantage of increasing the power delivered into arrays in preference to building large arrays is one requiring careful consideration of all the factors involved —the cost of land, masts and transmitters; maintenance charges; the life of valves; the cost of staffing; and the nature of the radio service. Obviously each specific case requires separate consideration. The widening of a beam can, of course, only be affected by sacrificing the ratio of field strength in the primary direction to the power input. In general, for mobile Services the best policy would appear to be to build several arrays, having different orientations but fairly narrow beams. The experiments of Wilmotte, to which reference is made by Mr. Barton, were valuable inasmuch as they were unique and gave an approximate picture of vertical polar diagrams. It is, however, debatable whether additional tests are worth while.

The feature of the Sterba array, mentioned by Mr. Gill, which enables sleet to be removed is valuable in countries having a more severe climate than England. As far as I am aware, Wilmotte first suggested* that it would be better to suppress more than a half a wave in

FIG. B.—Envelope of current distribution in radiating wire.

the Franklin type of aerial and adopt the type of current envelope shown in Fig. B. In some types of aerial this idea is now effected by folding the wire of the aerial in various ways. Mr. Gill's suggestion possibly offers another means of achieving the same result. In effect, instead of suppressing the radiation from the end of the half-wave elements, he proposed to allow the ends of adjacent half-wave radiators to overlap for the purpose of augmenting the radiation. The idea is ingenious but, without practical test or theoretical analysis, I am unable to state whether improvement would result. The action of one radiator on an overlapping radiator might unfavourably effect the arrangement.

Mr. Gill raises the question of dead-loss resistance of half-wave elements. I confess that 25 ohms is more in the nature of an intelligent guess than an exact measurement. The value will vary according to the location of the array; the type of towers and their distance from the array; the value of the insulation of the array; and the arrangement of wires in the array. Eckersley† gave 14 ohms as the measured value of the dead-loss resistance of a quarter-wave aerial on a 300-metre wavelength. According to Mr. Barfield, who has kindly given me permission to quote the results of some tests made by him, the dead-loss resistance of a single-wire vertical aerial 6 metres high at 25 metres wavelength

* See Bibliography, (3). † *Ibid.* (6).

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(approx.) was 31 ohms. This value was obtained by measuring the total resistance of the aerial and deducting the calculated value of the radiation resistance. Other evidence from a reputable source seemed to me to justify the assumption of 25 ohms for the purpose of a comparison of different types of aerial. The effect of an increase or decrease in the value of r will, of course, depend upon the values of R_0 and R_N in the expression $N(R_0 + r)/(R_N + r)$. With regard to transposing lines, tests made on open-wire lines appeared to lead to the conclusion that little was to be gained thereby. Each transposition point produces in the line constants a change which may be sufficient to nullify any advantage. Rotation of the wires in each limb would remove the possible objection of variation of impedance. With regard to the accuracy of the figures for efficiency of the open 2-wire transmission lines, I believe that, provided the out-of-balance currents are not too great, the figures are reasonably reliable. Mr. Gill mentions some of his tests on transmission lines. It is understood that the tests referred to were made on concentric tubes. An examination of the method of test and of the results obtained has convinced me that Mr. Gill's experiences cannot be advanced in refutation of the results on open transmission lines, quoted in the present paper. With regard to the value of Z_0, reference to the Section on transmission lines will show that this quantity cancels out in calculations of efficiency.

The shifting of the position of the current nodes mentioned by Squadron-Leader de Burgh is a common experience. In the case quoted the cause is probably lack of electrical symmetry in the array, owing to the presence of the mast stay-wires. As far as I am aware, there is no reliable field apparatus for measuring line impedances at radio frequencies of the order of 2×10^7 cycles.

The assumption of equal currents in the various radiators, to which reference is made by Mr. Carter, seems to be necessary for comparison purposes, particularly as the ratio of E to $\sqrt{P}$ is not greatly affected by appreciable variations in the relative current values in different radiators. For example, suppose an 8-wire vertical type of array has the following current values in adjacent wires, counting from one end and taking arbitrary units—1 : 2 : 3 : 4 : 4 : 3 : 2 : 1. The ratio $E/\sqrt{P}$ may be represented by $\Sigma I/\sqrt{[\Sigma I^2(R + r)]}$. If now the power is evenly divided between the 8 wires, the current values in each wire will be represented by 2·74 (approx.) and the ratio of field strength at a distant point due to the two systems of current distribution will be 20/21·9 for equal power input. Thus ordinarily the losses due to uneven current distribution need not be great. In the foregoing discussion the assumption is made, though it is not entirely justifiable, that all radiators have the same resistance. Mr. Carter's assumption that the parallel arrangement of radiators in the Koomans type of array produces more even current distribution, is hardly justifiable. The horizontal radiators, built one above the other, branch from the same 2-wire vertical feeder. Obviously, therefore, to obtain equal currents correct matching should be installed at each feeding point. Owing to the reaction of one radiator on another, quarter-wave spacings would not greatly affect the gains. Moreover, since Mr. Carter has already stated that a horizontal Bruce type of array has a very uneven current distribution in its members, and has conjectured that such an unequal distribution in radiators arranged one above the other would be disadvantageous, his reasons for suggesting the erection of a Bruce array turned through 90° and so building vertical tiers of radiators are not apparent. The tests made by Mr. Carter on the reflector action of a single wire placed behind a second single exciter wire are interesting and, in a measure, constitute a reply to Dr. Smith-Rose. It is understood, however, that the reflector wire was maintained at the same length throughout the tests. The results would have been more convincing if various lengths had been used for each distance of the reflector wire from the exciter wire. Mr. Carter raises the question of the best width of a beam in a vertical plane for reception purposes. The subject is very controversial, and in the absence of exact knowledge of the various factors involved any opinion on the matter is merely conjectural. For the purpose of the arguments of this paper, Fig. 13, having spans shown as horizontal ordinates, seems to me to be the best arrangement.

I note the claim of Mr. Chireix respecting his patent, but am unable to comment on this matter. With regard to the reduced cost of a multi-wire and concentric copper tube line for transmission purposes, the ratio of 5 : 1 was based upon actual costs and certainly was not biased in favour of the multi-wire line. For reception purposes small-diameter copper tubes can be used, and in this case the ratio of costs is approximately unity. I am interested to notice the confirmation given by Mr. Chireix of the critical effect observed with reflector wires.

Captain Eckersley refers to some of the many factors involved in the economic design of a beam radio station. A few of the factors have already been mentioned in the paper, and obviously others would have to be considered when the design of a new station is involved. The problem is simplified when development only of an existing station is proposed. In England, the economics applicable to the design of a large new station are influenced by two facts—the number of available suitable sites is very limited, and the cost of land-line connections quickly decides the economic distance of the site from the control terminal office. Obviously, it is uneconomical to have the site far removed from main trunk lines, so that existing circumstances largely decide the location of the radio station. When, however, the question of the power required to be radiated in a desired direction is considered, circumstance is not so helpful. There are sometimes long periods during which satisfactory service can be maintained on relatively small power; there are frequent occasions when a manifold increase in power would not convert an uncommercial into a commercial service. Thus the decision upon the correct power to be radiated, and hence the best ratio of expenditure on arrays and transmitting plant, is fraught with difficulties almost beyond solution. Regarding the question of maintenance of aerials, low aerials have obvious advantages. So far as the Post Office aerials are concerned, the principle of thoroughness is observed in construction, and in consequence maintenance charges

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due to broken wires and insulators are negligible. Moreover, the provision of a large telescopic ladder enables the greater part of the array described in the paper to be reached and thus efficiently maintained both as regards cost and time.

I was very gratified to see the communication of Mr. Franklin, whose pioneer efforts in short-wave beam working are remembered with admiration. Mr. Franklin quite correctly points out that there are limitations to the improvement in gain obtained by spacing one curtain of exciters half a wave behind the other. There is a slight narrowing effect both on the horizontal and vertical polar diagrams, but consideration of polar diagrams alone is equivalent to the assumption of 100 per cent radiation efficiency of the array. Dead-loss resistance, however, is appreciable and the higher this loss the nearer the field strength gain is simply proportional to the square root of the total number of radiators. I disagree with Mr. Franklin's contention that a more useful comparison of arrays could be made between the C type array having twice the height of curtain shown in Fig. 13, and the D type. The paper dwells upon the high cost of tall masts, and I definitely hold the view, both from theoretical and practical considerations, that the economic limit of height is quickly reached. Mr. Franklin refers to his scheme of folding wires to suppress radiation; the method which I described, whereby the current induced in a separate insulated wire is used to suppress unwanted radiation has been known for a long period and is believed to have originated in the Post office. As pointed out in the paper, the horizontal type of array has practically no side radiation, and suppression wires are therefore not required.

In reply to Mr. Franklin's general criticism, I can only assert that tests taken with local and distant transmissors show that the addition of a second exciter curtain used in the " T.W." type of array, is responsible for an additional improvement of about 2 decibels The account of the reasons responsible for the adoption of the concentric-tube type of feeder is interesting. Circumstances seem to have given Mr. Franklin no option in the matter. The simple truth, of course, is that the type of aerial employed, and shown diagrammatically in Fig. 2, is productive of large out-of-balance currents in open 2-wire transmission lines. As one line conductor is connected to the bottom of the aerial and the other to earth, obviously radiation from the open transmission lines would be very serious. The Post Office Engineering Department, realizing the value of symmetrically disposed radiators in an array, pinned their faith on open transmission lines. Results have entirely justified this policy. At a very small fraction of the cost of the concentric-tube line, high-efficiency lines have been built, possessing the advantages of easy and low maintenance charges.

Mr. Franklin states that with his type of aerial, overhead lines give rise to instability. This contrasts very unfavourably with the Rugby system. The fact is that the 16- and 33-metre services at this station using arrays fed through open lines 3 000 and 3 500 ft. long, respectively, with a feeder loss of about 30 per cent, are giving a service the frequency stability of which is unequalled. This must constitute my reply to Mr. Franklin's advocacy of concentric-tube lines.

Mr. Green draws attention to the apparent inconsistency in my reference to the reduction in resistance loss when three twin lines were used. In his arguments Mr. Green assumes similar lines, correctly terminated and having a surge impedance one-third that of a single twin line. All these assumptions are erroneous and therefore his conclusions fail. In his comments upon multi-wire lines, Mr. Green appears to have confused two factors—the relative ohmic loss of transmission lines having different surge impedances and the reduction in resistance loss obtained by dividing the copper in the limbs of a line whilst maintaining a definite surge impedance. The fact that low-impedance lines have greater ohmic losses than similar high-impedance lines for the same copper content was mentioned at the top of page 28 and hardly needs stressing. I did not suggest that multi-wire lines should be used as a substitute for 2-wire lines when radiation losses are low; I stated that multi-wire lines form an economical substitute for tube lines. The multi-wire lines can, of course, be arranged either in concentric or parallel lines. The point which I wished to bring out was that multi-wire lines built fairly close to the ground seem to reduce the radiation losses arising from unsymmetrical types of arrays. The possibility is that the unbalanced current, which must otherwise take a path via earth and transmitter capacitance, can, to a greater extent than is possible with single-wire lines, take the larger capacitance path between earth and line. In short, the multi-wire line tends to remedy the defect of unbalanced current due to unsymmetrically disposed radiators in an array.

Regarding the concluding remarks of Mr. Green, it is perhaps relevant to point out that telephone engineers are not unfamiliar with transmission-line problems and that much practical and theoretical work, both published and unpublished, has been undertaken since Oliver Heaviside first published his mathematical investigations into cable transmission problems. With regard to the comparatively simple application to radio engineering, correct scientific principles have been applied in general to problems of connection to aerial systems since Messrs. Beverage, Kellogg and Rice published the results of their theoretical and practical investigations in 1923.

I should like to express my sincere thanks for the very cordial reception given to my paper.

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A GRAPHICAL METHOD FOR DETERMINING THE MAGNITUDE AND PHASE OF THE ELECTRIC FIELD IN THE NEIGHBOURHOOD OF AN ANTENNA CARRYING A KNOWN DISTRIBUTION OF CURRENT.*

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[From the National Physical Laboratory.]

(*Paper first received 4th September, and in final form 12th November*, 1930.)

SUMMARY.

The development of the use of short waves and modern beam systems has given rise to a need for a knowledge of the electric field in the immediate neighbourhood of a transmitting aerial. The mathematical calculation of this field is somewhat laborious, and the following paper describes a graphical method of procedure by means of which calculations may be made for any desired arrangement of a transmitting aerial.

A simplified form of Hertz's equations is used to give the field at a point on the ground due to a unit doublet at different heights in a vertical antenna two wavelengths long. A curve is drawn connecting the amplitude of the field at the point with the height of the doublet in the antenna. For convenience of integration of this field along the antenna, the amplitude is divided into two components—one in phase with the current in the doublet and the other in quadrature with it. Curves are given showing the variation of these two components of the electric field with height for distances of the point from the base of the aerial of 0·25, 0·5, 0·75 and 1·0 times the wavelength of the current in the doublet. The field at any intermediate distance can be found by interpolation. The unit doublet used in the calculation of these curves is supposed to carry a current of 1 ampere. The field due to any small element of the aerial with another current distribution is equal to that when it carries 1 ampere multiplied by the current in amperes now in the element. If the two components of the electric field at the point considered due to unit current are modified in this way for a few discrete points in the antenna, two new curves may be drawn connecting the two components of the electric field due to any element in the antenna and the height of that element. The areas which these two curves make with the axis denoting height give the total electric fields at the point due to the complete antenna, and in phase and in quadrature respectively with the current in it. The magnitude of the resultant field is equal to the square root of the sum of the squares of the two components, and its phase is given by the inverse tangent of their ratio. The phase is determined throughout with respect to the current at the base of the antenna. It is then shown that the method may be applied to find the field at points other than those on the radius vector through the base normal to the antenna. In this way the field may be determined at various points in another parallel antenna.

The phase of the field is determined by the same method at various distances from different lengths of antenna, and a figure is given which shows that for distances greater than about $0 \cdot 3\lambda$ there is a maximum difference of about 5° between the phase of the field of a half-wave antenna and that due to the doublet at its base. This small difference explains why certain calculations based on the assumption that an antenna is small compared with the wavelength are approximately true when applied to antennæ in which this assumption no longer holds.

The effect of the earth is next taken into account and it is shown how the same method may be used to find the field at any distance from an antenna when the dielectric constant and the conductivity of the earth are known.

* Reprinted from *Journal I.E.E.*, 1931, vol. 69, p. 290.

(1) INTRODUCTION.

The discovery that medium and short radio waves can be reflected by the Heaviside layer has given an enormous impetus towards the use of antenna beam systems for point-to-point radio communication. These antenna systems should be designed to direct a beam of radiation at suitable angles of incidence to the Heaviside layer and, by this means, to reduce the losses in short-wave propagation due to absorption in the earth. Before the advent of the beam system problems on wave propagation consisted usually in finding the field at great distances from a simple vertical antenna the dimensions of which were so small compared with the wavelength that it could be considered as a Hertzian doublet. In order to produce a beam, however, the dimensions of the antenna system must be comparable with the wavelength, and its design necessitates a knowledge of the reactions between the various elements of the system. In this problem we can no longer consider the dimensions of the antenna array, nor even its components, as negligible, as they may be several times the distance between the various elements the mutual reactions of which we wish to study.

Pierce† has developed a method, first suggested by Abraham,‡ for determining the radiation characteristics of an antenna by dividing it into a series of Hertzian doublets and integrating the field due to each at a given point. The integrals be obtains, however, are too complicated for use in any practical problem at short distances, and a graphical method of integration has, therefore, been developed in the present paper. A solution by Pierce's method can only be obtained if the distribution of current in the antenna conforms to some simple mathematical function. This limitation does not apply to the graphical method to be described, which feature makes it particularly suitable for analysing the mutual effects of the components of antenna arrays.

The effect of an electromotive force in producing

† See Reference (1). ‡ *Ibid.* (2).

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current in an antenna depends upon its point of application. Any problem on the propagation of waves from an antenna array can be solved, once a knowledge of the current in each element is known. The method described

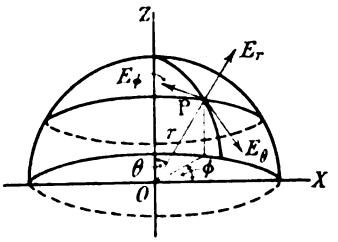

FIG. 1.

below shows how the e.m.f. at all points of an array may be found when the dielectric constant and conductivity of the ground are known. If the factor which determines the ratio of current to e.m.f. is obtained at a

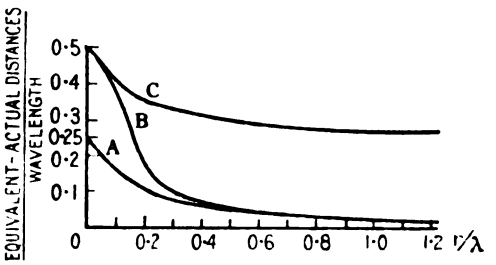

FIG. 2.

Curve A: $y = (s_\phi - r)/\lambda$
Curve B: $y = \frac{1}{2}[(r_\theta - r)/\lambda] - \frac{1}{2}$
Curve C: $y = (r_r - r)/\lambda$

few discrete points, the current in all parts of the array may be found by interpolation and, hence, the field strength at any point outside the array determined. In the present paper the problem is limited to finding the

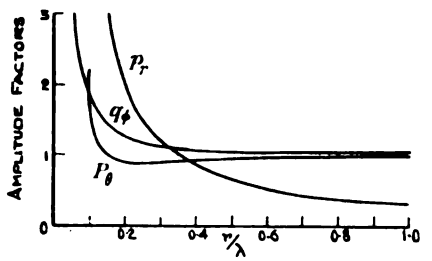

FIG. 3.

e.m.f. at different distances from an antenna carrying a known distribution of current.

(2) FIELD DUE TO DOUBLET.

Wilmotte and McPetrie* have shown that the Hertzian equations of the field at any point (r, θ, ϕ), Fig. 1, due to any sinusoidal component $f(t)$ of an element of current-length flowing in the direction $\theta = 0$ at the origin can be expressed in the form

$$H_r = H_\theta = E_\phi = 0 \quad . \quad . \quad . \quad . \quad (1)$$

* See Reference (3).

$$E_\theta = \frac{30p_\theta}{cr} \sin\theta f'\left(t - \frac{r_\theta}{c}\right) \text{ volts/cm} * \quad . \quad . \quad (2)$$

$$E_r = \frac{30p_r}{cr} \cos\theta f'\left(t - \frac{r_r}{c}\right) \text{ volts/cm} \quad . \quad . \quad (3)$$

$$H_\phi = \frac{30q_\phi}{cr} \sin\theta f'\left(t - \frac{s_\phi}{c}\right) \text{ volts/cm} \quad . \quad . \quad (4)$$

in which

$$r_\theta - r = \frac{\lambda}{2} + \frac{\lambda}{2\pi} \arctan \frac{2\pi(r/\lambda)}{4\pi^2(r/\lambda)^2 - 1} \quad . \quad (5)$$

$$r_r - r = \frac{\lambda}{2} - \frac{\lambda}{2\pi} \arctan \frac{2\pi r}{\lambda} \quad . \quad . \quad . \quad (6)$$

$$s_\phi - r = \frac{\lambda}{2\pi} \arctan \frac{\lambda}{2\pi r} \quad . \quad . \quad . \quad (7)$$

$$p_\theta = \sqrt{\left[1 - \frac{\lambda^2}{4\pi^2 r^2} + \frac{\lambda^4}{16\pi^4 r^4}\right]} \quad . \quad . \quad (8)$$

$$p_r = \frac{\lambda}{\pi r}\sqrt{\left[1 + \frac{\lambda^2}{4\pi^2 r^2}\right]} \quad . \quad . \quad . \quad . \quad (9)$$

and

$$q_\phi = \sqrt{\left[1 + \frac{\lambda^2}{4\pi^2 r^2}\right]} \quad . \quad . \quad . \quad . \quad (10)$$

in which c is the velocity of light and c/λ the frequency of $f(t)$.

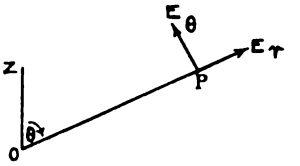

FIG. 4.

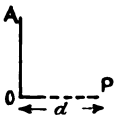

FIG. 5.

In the above-mentioned paper the factors r_θ, r_r and s_ϕ were called "equivalent distances," and p_θ, p_r and q_ϕ "amplitude factors." The variation with distance of these quantities is shown in Figs. 2 and 3. The use of these curves makes it an easy matter to deduce from equations (2), (3) and (4) the phase and magnitude of

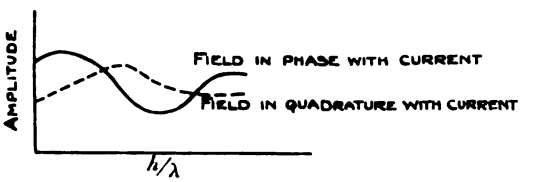

FIG. 6.

the electric and magnetic fields at any distance due to a given element of current-length at the origin. It should be noted that the positive direction of E_θ in equation (2) is in the direction of decreasing θ. This means that along the horizontal line normal to the doublet $(\theta = \frac{1}{2}\pi)$ the positive direction of E_θ is vertically upwards.

Let O, Fig. 4, represent the position of the current-element $f(t)$ the direction of which is $OZ(\theta = 0)$. It may be seen from the above equations that at any distance E_θ leads $f(t)$ in phase by an angle $\frac{\pi}{2} - \frac{2\pi r_\theta}{\lambda}$.

* In equations (2), (3) and (4) the current $f(t)$ should be expressed in amperes and all distances in centimetres.

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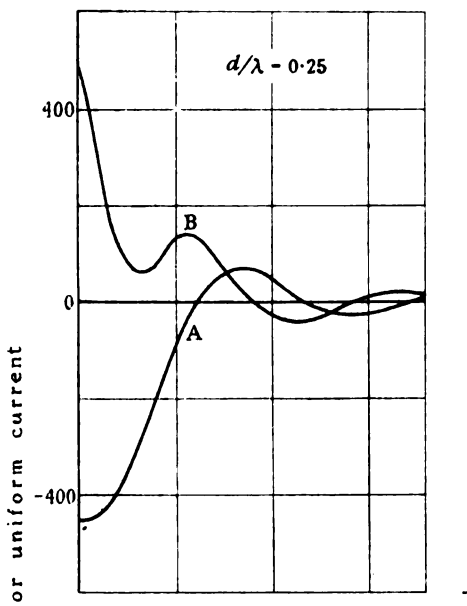

FIG. 7.

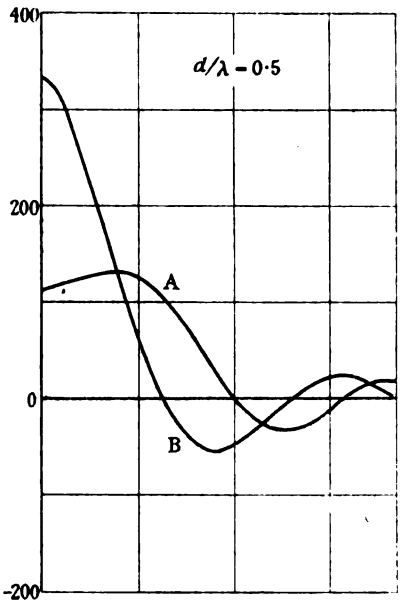

FIG. 8.

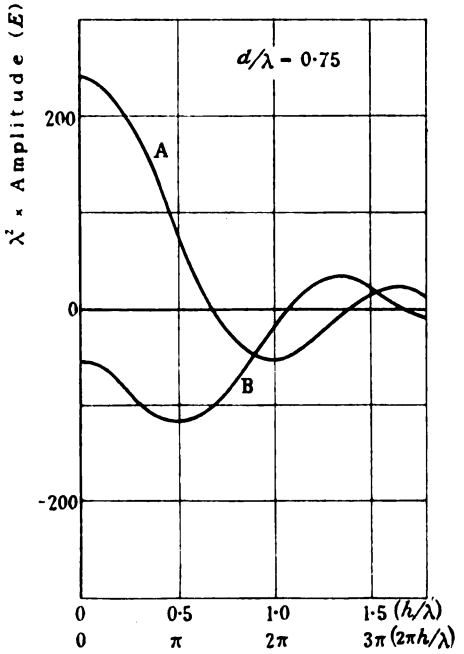

FIG. 9.

Curves A:—Vertical electric field in phase with current.

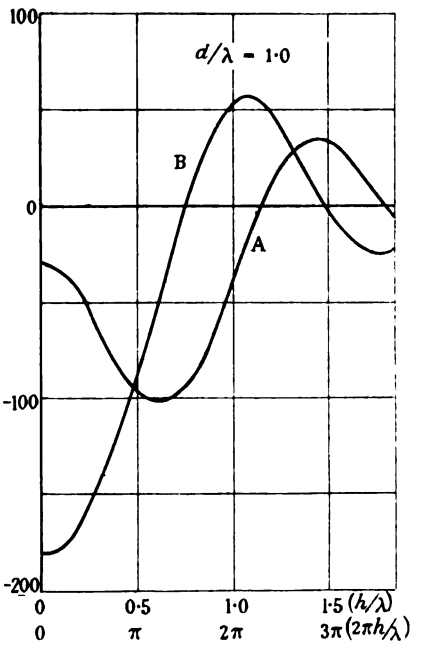

FIG. 10.

Curves B:—Vertical electric field in quadrature with current.

Similarly E_r leads $f(t)$ by $\frac{\pi}{2} - \frac{2\pi r_r}{\lambda}$. The total electric field at $P(r, \theta, \phi)$ therefore, which is parallel to the positive direction of OZ and in phase with the current $f(t)$, is

$$\left\{\frac{60\pi p_\theta}{\lambda r}\sin^2\theta\cos\left(\frac{\pi}{2}-\frac{2\pi r_\theta}{\lambda}\right) + \frac{60\pi p_r}{\lambda r}\cos^2\theta\cos\left(\frac{\pi}{2}-\frac{2\pi r_r}{\lambda}\right)\right\}f(t) \quad (11)$$

and the total field in the same direction but in quadrature with and leading $f(t)$ is

$$\left\{\frac{60\pi p_\theta}{\lambda r}\sin^2\theta\sin\left(\frac{\pi}{2}-\frac{2\pi r_\theta}{\lambda}\right) + \frac{60\pi p_r}{\lambda r}\cos^2\theta\sin\left(\frac{\pi}{2}-\frac{2\pi r_r}{\lambda}\right)\right\}f(t) \quad (12)$$

(3) Field at Point Due to Uniform Current in Aerial.

Let OA, Fig. 5, represent an antenna carrying a uniform distribution of current. This condition is not

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realizable in practice but it will be shown later how the results obtained may be modified to take account of any given current distribution in OA. Neglecting the effect of neighbouring conductors and also the image of OA in the earth, we can, from equations (11) and (12), calculate the electric field parallel to OA at any point due to any doublet in OA. We shall limit ourselves, at present, to finding the field at points on the radius vector through O perpendicular to OA. Let P, Fig. 5, represent one of these points distant d from O. We can consider the aerial OA as consisting of an infinite number of doublets, and integrating vectorially the fields at P due to these elements we find the total field at P due to the given current in OA. In order that this integration may be performed graphically we shall calculate the components of these elementary fields which are in phase and in quadrature with the current in the transmitting aerial OA. By adding the same components of these fields, we obtain the total fields due to OA at P which are in phase and in quadrature with the current in OA.

Suppose these elementary fields at P are obtained for a number of equal doublets at different heights in OA and that curves are drawn as in Fig. 6, in which the amplitudes of the fields at P parallel to OA and in phase and in quadrature with the current in it are plotted as ordinates against h/λ, where h is the distance from O of the doublet in OA and λ is the wavelength of the current in OA. The fields at P due to the complete aerial OA which are parallel to OA and in phase and in quadrature with the current in it are proportional to the areas which the respective curves in Fig. 6 make with the h/λ-axis. The total field at P due to the complete aerial OA is, therefore, fully determined as its magnitude is equal to the square root of the sum of the squares of the two fields, and its phase with respect to the current in OA is given by the inverse tangent of their ratio. The constant of proportion between the areas measured and the field at P may easily be determined in any required units. The equations given above, which apply only to the radiation from an infinitesimal doublet, may therefore be applied to find the field at any point in space due to uniform current in a linear antenna the dimensions of which are no longer infinitesimal compared with the wavelength of the current in it. The distribution of current in a radiating antenna can never be uniform and we shall only use the uniform-current case as a basis for finding the field at a point due to any non-uniform distribution of current in a given antenna.

Figs. 2 and 3, along with equations (11) and (12), have been used to obtain Figs. 7, 8, 9 and 10, in which the components of the amplitude of the vertical electric field at P (Fig. 5) in phase and in quadrature with unit doublet in OA, are plotted against the height of the doublet in OA. The figures correspond to distances of OP of $0{\cdot}25\lambda$, $0{\cdot}5\lambda$, $0{\cdot}75\lambda$ and λ respectively. The ordinates are expressed in volts multiplied by the square of the wavelength when the current at each point in OA is assumed to be 1 ampere; and in order that the curves may be used for any wavelength all distances are expressed in terms of the wavelength. The field due to unit doublet at a given height in OA and at a distance intermediate between any of those given in the above figures may be found by interpolation.

(4) Field at Point Due to Aerial Carrying Non-Uniform Current.

We have considered the current in OA (Fig. 5) to be uniform. This condition is not realizable in practice. We shall now show how Figs. 7, 8, 9 and 10 may be used to give the field at the same distances when the current in OA has any given distribution. Suppose the amplitude of the current in OA is of the form shown by the solid line in Fig. 11 and that the broken straight line represents uniform current of 1 ampere on the same scale. The field at P due to the element δl in OA has been found when the current throughout its length is 1 ampere. The field at P due to the same element when the current in it is as shown in Fig. 11 is equal to the field at P for 1 ampere in δl multiplied by the ratio of the two areas "abcd" and "abef." The areas of these two figures may be calculated or found by planimeter. By finding the ratio of these two areas for different positions of δl in OA, we obtain different constants with which to multiply the ordinates of Figs. 7, 8, 9 and 10 for the corresponding values of h/λ. A new series of curves is

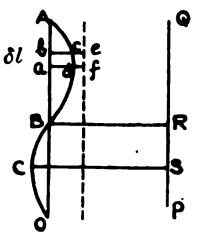

Fig. 11.

then obtained in which the amplitude of the field at P, due to the doublet δl in OA, is plotted against h/λ, where h is the distance of δl from O. The areas which these new curves make with the h/λ-axis determine the field at P due to the given distribution of current in OA. Hence, once we have obtained the field at various distances from an antenna carrying uniform current we can use the results to give the field at the same distances from an antenna with any given distribution of current.

Figs. 12, 13, 14 and 15 give the fields in phase and in quadrature with the current in OA at various distances along a line normal to the antenna through one end.* The aerial is assumed to be 1 wavelength long and to carry a sinusoidal distribution of current, the nodes being at the centre and the two ends. When the magnitude of the current at a loop is 1 ampere, the unit of amplitude is in volts multiplied by the square of the wavelength, the latter being expressed in centimetres. For any other magnitude of current at a loop the amplitude should be multiplied throughout by that current expressed in amperes. The total field at P, due to any part of the antenna OA, is given directly by the areas the curves make with that part of the h/λ-axis which corresponds to the portion of OA under consideration. The field at P, for instance, due to the lower loop of current in OA, is given by the areas of

* The phase is always determined with respect to that of the current immediately above the base of OA.

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the curves when the limiting values of h/λ are 0 and 0·5, and that due to the upper loop when the limits of h/λ are 0·5 and 1.

One set of curves, therefore, gives us the field due to any length of antenna up to 1 wavelength in height in through a loop of current in OA. The units are the same as those in Figs. 12 to 15, and the field due to the various parts of the antenna may be found, as shown above, by measuring the areas of the curves between the appropriate values of h/λ.

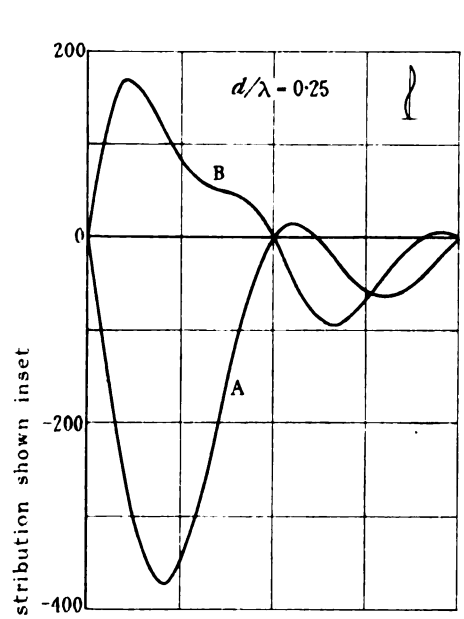

FIG. 12.

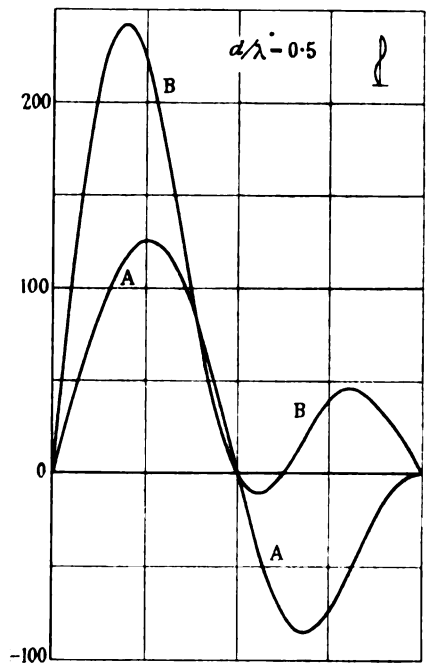

FIG. 13.

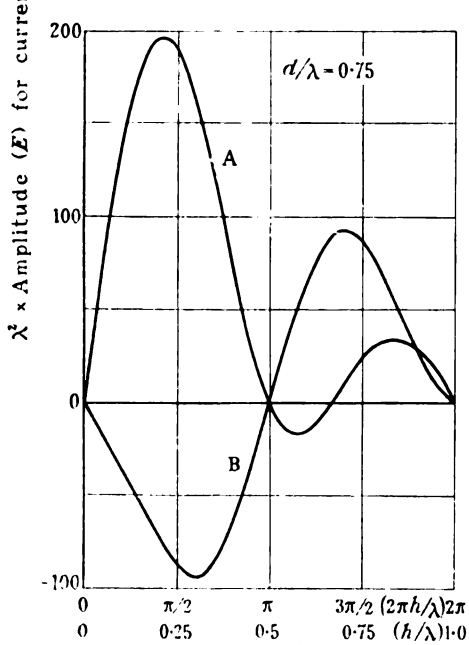

FIG. 14.

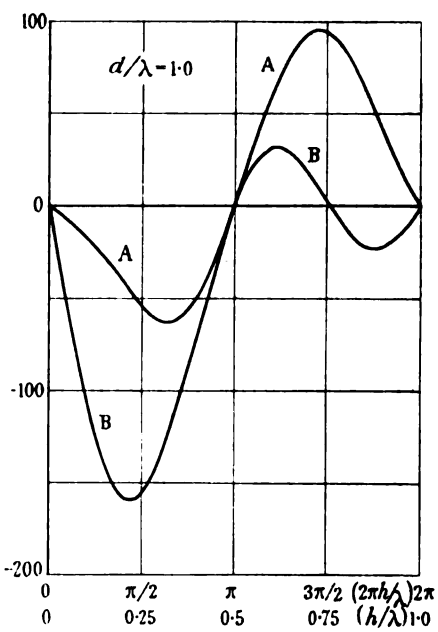

FIG. 15.

Curves A :—Vertical electric field in phase with current at base of aerial.
Curves B :—Vertical electric field in quadrature with current at base of aerial.

which the current is sinusoidal and zero at the point on the antenna opposite that at which the field is considered. Figs. 16, 17, 18 and 19 give the fields at the same distances when the point P is on the radius vector

(5) METHOD OF FINDING FIELD AT VARIOUS POINTS IN A WIRE PARALLEL TO A TRANSMITTING ANTENNA.

We shall now show how the above curves may be applied to find the field at various points in a second

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antenna parallel to OA, if OA carries, let us say, the current shown in Fig. 11. Suppose the distance OP is equal to $0 \cdot 25\lambda$. The in-phase and in-quadrature fields at P are given directly by the areas of the complete line BR. The field at S opposite the lower loop of current in OA is the same as that due to an aerial with current distribution, the same as in the part AC of AO along with one having the distribution as in CO. The field at S

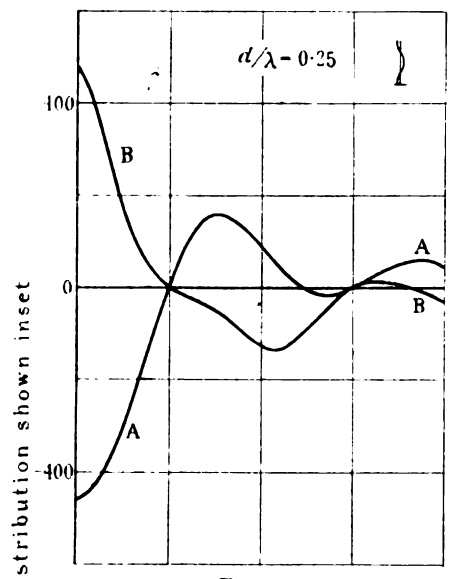

FIG. 16.

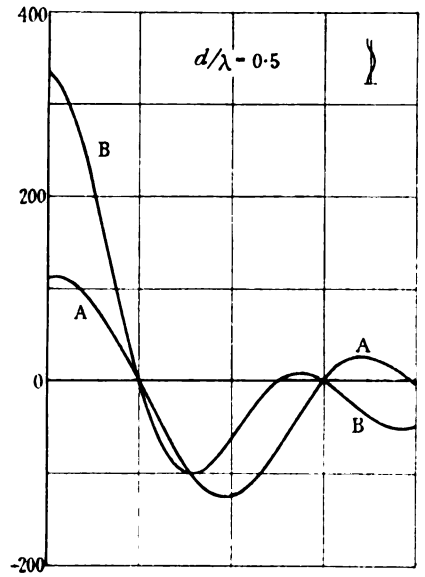

FIG. 17.

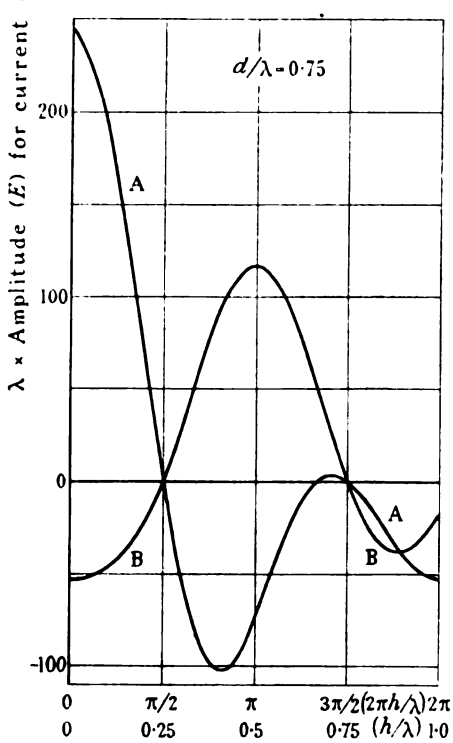

FIG. 18.

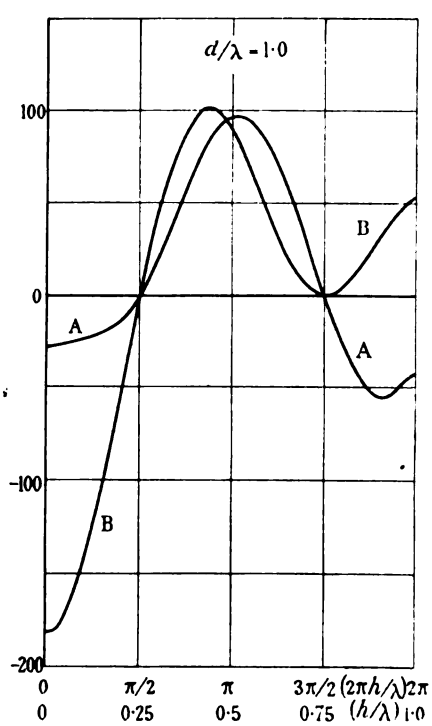

FIG. 19.

curves in Fig. 12. The vertical field at Q is equal in magnitude to that at P but opposite in phase, due to the reversal in phase of the current in the upper loop of OA. The field at R, the midpoint of P, is zero as the two loops of current in OA neutralize one another along the due to the part AC of AO is given by the areas of the curves in Fig. 16, when the limits of h/λ are 0 and $0 \cdot 75$. The field due to CO is given by the areas of the same curves when h/λ is limited between 0 and $0 \cdot 25$. By adding the two fields in phase and in quadrature with

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the current in the lower loop of AO, we obtain the total in-phase and in-quadrature fields at S due to the aerial AO having the given current distribution. Thus the magnitude and phase of field at S are completely determined. The field at T opposite the loop of current in the upper half of AO is equal to that at S but of opposite phase. From Figs. 12 and 16, therefore, we have completely determined the field at 5 points on an antenna 1 wavelength long, distant $0{\cdot}25\lambda$ from an excited antenna of the same dimensions. This particular case has only been given as an example of the method by which Figs. 12 to 19 may be applied with ease to find the field at various points and distances from an antenna having any given current distribution. It may be considered that the points on the second aerial at which the field is determined are too far apart. More points may be considered if we use Figs. 7 to 10 to find the field at points opposite not only loops and nodes of current but also various parts of a sinusoidal current distribution.

(6) Total Field Induced in an Antenna by Current in Another.

Burnett* relates the current I to the electromotive force E in an antenna having inductance and capacitance per unit length of L and C respectively by the equation

$$\frac{\partial^2 I}{\partial h^2} + \omega^2 LCI = -i\omega CE \quad . \quad . \quad . \quad (13)$$

where h denotes the distance from one end of the point at which E and I are measured. This equation shows that the effect of a given e.m.f. depends on its point of application. This means that after finding the e.m.f. at a given point in an antenna we must reduce it to what may be termed an effective e.m.f. at that point. The solution of the above differential equation is too complicated for finding this reduction factor, but it is hoped that a simple approximation may soon be obtained. If such a reduction factor is found for various points on an antenna it will be possible to integrate the effective e.m.f. along the aerial. This may be done by measuring the area which a curve of effective e.m.f. makes with the axis of h. Such an integration would give the total field induced in the antenna. The method described above cannot, therefore, determine the total field induced in one antenna by another, but no method has been given by which the field at various points on this second antenna can be so rapidly determined.

* See Reference (4).

(7) Phase of Field at Various Distances from Antenna.

The solid lines in Fig. 20 give the variation of the angle by which the phase of the vertical electric field from a doublet leads the current in that doublet for different distances along the line normal to the doublet. The dotted lines were obtained by the method described in this paper, and give the phase of the field for various distances along the normal through one end of a $\frac{1}{2}$-wavelength antenna. It will be noticed that, except at small distances, there is a maximum difference of about 5° between the two sets of curves. The difference between the doublet and antennæ smaller than $\frac{1}{2}$ wavelength is less than this 5°. This small difference explains why certain calculations based on the assumption that the antenna is infinitesimal compared with the wavelength are approximately correct when applied to antenna arrays in which the components are no longer small compared with the wavelength. Now, consider, say, two $\frac{1}{2}$-wavelength antennæ a given distance apart, one antenna being excited directly by a transmitter. The phase of the field at each end of the unexcited antenna relative to the current in the excited one is given by the

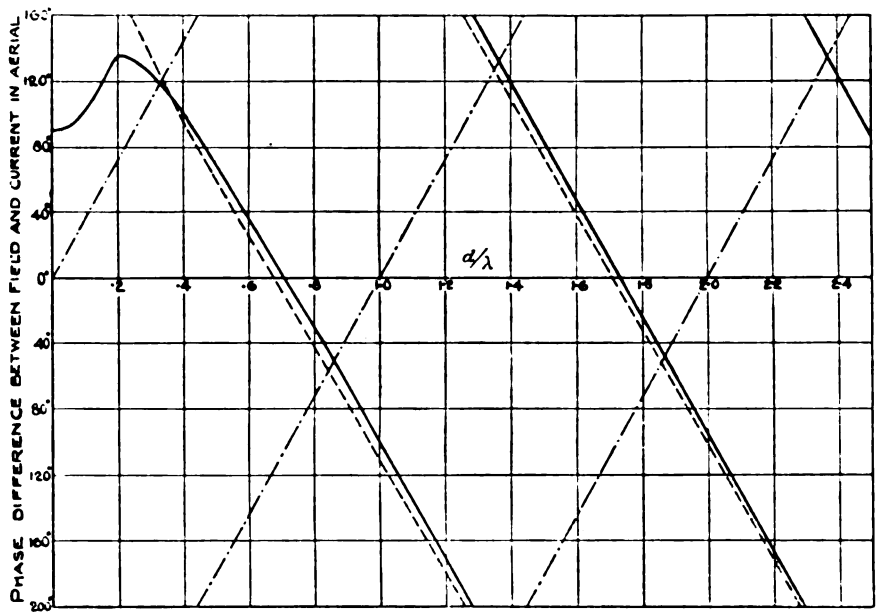

FIG. 20.

—— Doublet. - - - Half-wave antenna. — · — · $2\pi d/\lambda$.

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dotted lines in Fig. 20. The field at the centre of the unexcited antenna which is opposite a loop of current in the primary antenna may be considered as due to two $\frac{1}{4}\lambda$-antennæ at the excited antenna which are joined together. The phase of the field at a distance from a $\frac{1}{4}\lambda$-antenna, as shown above, differs from that of a $\frac{1}{2}\lambda$-antenna by less than 5°. The phase of the field, therefore, at the centre of the secondary antenna differs from that at its ends by less than 5°. The field at other points on the second antenna differs by less than this and so we have the result that the phase along the second antenna differs from that at its ends by less than 5°. This, of course, is true only if the distance between the two antennæ is such that the two curves in

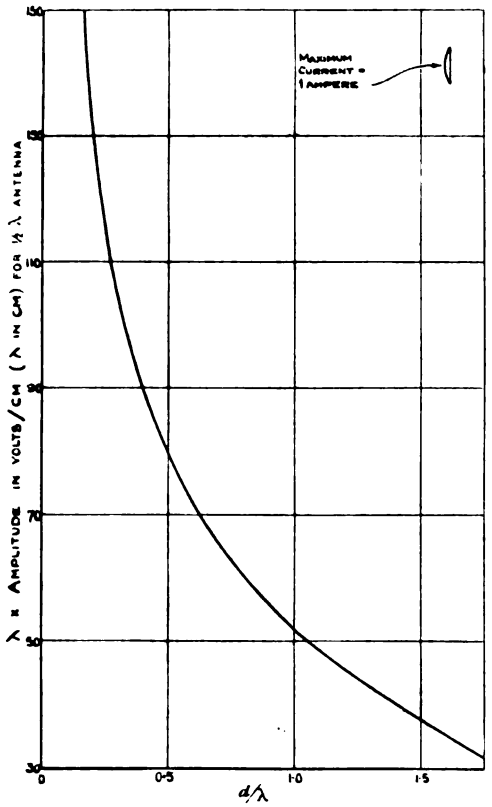

FIG. 21.

Fig. 20 are in close proximity, that is, for distances greater than about $0{\cdot}3\lambda$. The phase of the field at different distances from a doublet may be found directly from Hertz's equations. It may also be found from the graphical method described above by assuming the length of the antenna to become very small. Both methods have been used and the closeness with which the results agree shows that the graphical method can determine the phase of the field at any point to within about 1°.

Another result may be obtained from Fig. 20. Suppose we have two aerials, A and B, one of which (B) is excited by induction from the other, and suppose the problem is to find the distance AB so that when B is tuned to the frequency of the current in A the electric fields from A and B are in phase at great distances in the direction BA. In order that this condition may be satisfied the current in B must lead the current in A by an angle equal to $2\pi l/\lambda$ times the distance AB. The chain lines in Fig. 20 are obtained by plotting this angle $2\pi d/\lambda$ against d/λ. The points of intersection of these lines with the solid and dotted curves, therefore, give the positions in which B must be placed in order to satisfy the required conditions. It will be seen that the smallest value of AB is about $0{\cdot}33\lambda$ and that its magnitude is practically the same whether the aerial A is a doublet or equal in length to $\frac{1}{2}\lambda$. Palmer and Honeyball* have found experimentally that when AB is $0{\cdot}33\lambda$ and B is tuned to the frequency of the current in A, there is maximum radiation in the direction BA.

(8) Variation of Field Strength with Distance from Antenna.

The variation of field strength with distance from a $\frac{1}{2}\lambda$-antenna is shown in Fig. 21. The field strength at the different distances was obtained by integrating the in-phase and in-quadrature field curves of Figs. 12 to 15 for the values of h/λ between 0 and $0{\cdot}5$. The total field at any point is given by the square root of the sum of the squares of its two components. Similar curves may be obtained giving the relation between field strength and distance for different types of antenna.

(9) Method of Taking Account of the Image of the Antenna in the Earth.

We have, so far, neglected the image of the antenna in the earth. A complete solution of any problem on the field in the neighbourhood of an antenna must take account of this image. The author will endeavour to show how the results obtained above may be modified so as to give the field at any point when the dielectric constant and the conductivity of the ground are known.

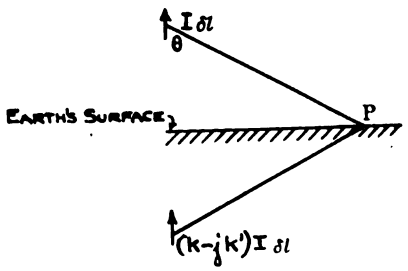

FIG. 22.

Wilmotte* shows that an element of current $I\delta l$ (Fig. 22) has for image a similar element but of strength $(K - jK')I\delta l$, where K and K' are functions of the angle of transmission to the vertical and depend on whether the element lies in or perpendicular to the plane in which the propagation is considered.

Throughout this paper we have considered only that part of the field in the plane of propagation. When the element of current lies in this plane K and K' are represented by K_i and K_i' and have values determined by the relation

$$K_i - jK_i' = \frac{(\epsilon^2 + 4\sigma^2/f^2)\cos^2\theta - (c^2 + d^2) - 2j\cos\theta(2c\sigma/f + \epsilon d)}{(\epsilon^2 + 4\sigma^2/f^2)\cos^2\theta + (c^2 + d^2) + 2\cos\theta(\epsilon c - 2d\sigma/f)}$$

* See Reference (5).

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in which j is equal to $\sqrt{-1}$, σ is the conductivity of the ground in electrostatic units, ϵ the dielectric constant, f the frequency, and c and d are given by

$$c^2 - d^2 = \epsilon^2 - \sin^2\theta \quad \text{and} \quad cd = -\sigma/f$$

Consider an element of current $I\delta l$, Fig. 22. The image in the ground is $(K_i - jK_i')I\delta l$. The vertical electric field at P due to $I\delta l$ can be obtained from the appropriate figure in the earlier part of this paper. To take complete account of the effect of the earth we have to multiply this elementary field by $(1 + K_i - jK_i')$, the values of K_i and K_i' being given by curves in the paper referred to above.* If this procedure is followed for various positions of the element of current, the total vertical electric field due to the element and its image at any point on the ground may be determined. Two new figures may now be drawn, giving respectively the field at the required point in phase and in quadrature with the current and due to elements at different heights in the antenna. The magnitude and phase of the field due to the complete antenna can then be determined in the way outlined above. The effect of the earth will be studied more fully in a later paper.

This work was carried out as part of the programme of the Radio Research Board, and the author is indebted to the Department of Scientific and Industrial Research for permission to publish this paper. He desires also to acknowledge his indebtedness to Dr. R. L. Smith-Rose for valuable advice on the method of presentation of the paper and to Mr. B. J. Byrne for assistance in the computing work involved.

* See Reference (6).

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(1) G. W. PIERCE: "Electric Oscillations and Electric Waves," 1920, chap. 9.

(2) M. ABRAHAM: "Spark Telegraphy and Electrodynamics," *Physikalische Zeitschrift*, 1901, vol. 2, p. 329.

(3) R. M. WILMOTTE and J. S. MCPETRIE: "A Theoretical Investigation of the Phase Relations in Beam Systems," *Journal I.E.E.*, 1928, vol. 66, p. 949.

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(5) L. S. PALMER and L. L. K. HONEYBALL: "The Action of a Reflecting Antenna," *Journal I.E.E.*, 1929, vol. 67, p. 1045.

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MEETINGS OF THE WIRELESS SECTION.

85TH MEETING OF THE WIRELESS SECTION, 5TH NOVEMBER, 1930.

Mr. Clifford C. Paterson, O.B.E., President, took the chair at 6 p.m.

The minutes of the meeting of the Wireless Section held on the 7th May, 1930, were taken as read and were confirmed and signed.

A vote of thanks to Captain C. E. Kennedy-Purvis for his services as Chairman of the Wireless Section during the session 1929–30, proposed by Commander J. A. Slee, C.B.E., R.N., and seconded by Mr. J. F. Herd, was carried with acclamation.

Mr. C. E. Rickard, O.B.E., Chairman of the Wireless Section, then delivered his Address (see page 1).

A vote of thanks to Mr. Rickard for his Address was proposed by the President and, after being put to the meeting, was carried with acclamation.

The meeting terminated at 7.35 p.m.

86TH MEETING OF THE WIRELESS SECTION, 3RD DECEMBER, 1930.

Mr. C. E. Rickard, O.B.E., Chairman of the Section, took the chair at 6 p.m.

The minutes of the meeting of the Wireless Section held on the 5th November, 1930, were taken as read and were confirmed and signed.

A paper by Mr. T. Walmsley, B.Sc., entitled "Beam Arrays and Transmission Lines" (see page 15), was read and discussed.

The meeting terminated at 7.50 p.m. with a vote of thanks to the author, which was moved by the Chairman and carried with acclamation.

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A NEW METHOD OF MEASUREMENT OF RESISTANCE AND REACTANCE AT RADIO FREQUENCIES.*

By F. M. COLEBROOK, B.Sc., and R. M. WILMOTTE, M.A., Associate Member.

(*Paper first received 22nd February, and in final form 29th November, 1930; read before the* WIRELESS SECTION *7th January, 1931.*)

SUMMARY.

A method of measuring the radio-frequency resistance of coils is described. It depends on the change in reactance of a valve-maintained oscillating circuit when it is coupled to the circuit under investigation, and does not require any measurement of current or voltage. Practical details of the working of the method and the construction of the necessary apparatus are given. It is further shown how the method can be applied to the measurement of unknown impedances (e.g. the input impedances of valves, the impedances of aerials, etc.) by associating these impedances in some suitable manner with a coil and measuring the apparent change in the constants of the coil. Measurements have been made at about 10 million cycles per second, but the method could probably be used at higher frequencies than this.

(1) INTRODUCTION.

The method is essentially a development of one originally devised by P. W. Willans, and described by P. K. Turner.† It depends on the measurement of frequency and capacity, so that a wavemeter and calibrated condensers of good quality are required, but no other measuring instruments are necessary. It is distinguished in this respect from the method of Mallet and Blumlein,‡ which it otherwise resembles, inasmuch as it depends on the effect produced by the coil under investigation on another tuned circuit to which it is inductively coupled.

A description of the Willans method will first be given in order to make clear the nature of the modifications embodied in the present proposals.

(2) THE WILLANS METHOD.

Circuits 2 and 3 (Fig. 1) represent valve-maintained radio-frequency oscillators. Circuit 2 being set at the required measurement frequency, circuit 3 is adjusted to beat with it at some suitable audible frequency. Circuit 1, which consists of the coil under investigation and a calibrated condenser, is then closed. As C_1 is varied the beat frequency will be heard to vary after the manner shown in Fig. 2, in consequence of a corresponding variation of the frequency of oscillation of circuit 2. The values of C_1' and C_1'' corresponding to the maximum and minimum beat frequencies are noted, and the resistance R_1 is calculated from the formula

$$R_1 = \frac{1}{2\omega}\left(\frac{1}{C_1'} - \frac{1}{C_1''}\right)$$

The derivation of the formula is given in Appendix I.

(3) LIMITATIONS OF THE WILLANS METHOD.

The method is open to criticism in the following respects:—

(*a*) In the derivation of the formula (see Appendix I) certain effects caused by the variation of ω are neglected.

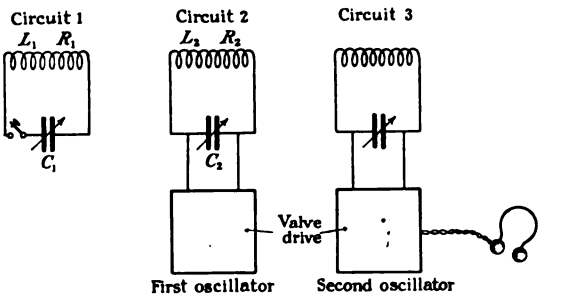

FIG. 1.

It is true that the error so introduced will be small in many practical cases, but in certain instances it may amount to as much as 100 per cent.

(*b*) It is difficult to determine the highest and lowest frequency points very exactly with the simple arrangement shown.

(*c*) For sensitivity the variation in ω should be large.

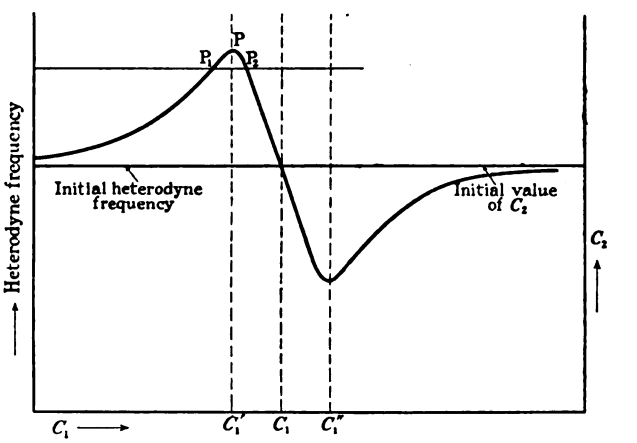

FIG. 2.

This variation can be controlled by alteration of M, i.e. by adjustment of the relative positions of circuits 1 and 2. This, however, will increase the errors mentioned in (*a*). Moreover, the variation in ω is limited in any case to the range of audible frequencies over which the ear is sensitive to small variations in pitch.

For these reasons the method lacks precision, though

* Reprinted from *Journal I.E.E.*, 1931, vol., 69, p. 497.
† *Experimental Wireless*, 1925, vol. 2, p. 350.
‡ "A New Method of High-Frequency Resistance Measurement," *Journal I.E.E.*, 1925, vol. 63, p. 397.

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its admirable simplicity will commend it for use in cases where high accuracy is not required.

(4) The New Method.

In the new method the above limitations are avoided as follows:—

(i) All measurements are made at the same frequency, $\omega/(2\pi)$.

(ii) The telephones carry an audio-frequency current of constant frequency supplied by an auxiliary audio-frequency oscillator. The beat frequency is matched with this constant audio-frequency by the usual method of beats. This affords an exceedingly sensitive method of synchronization.

The procedure is as follows:—(i) Open the switch in circuit 1. Adjust the frequency of circuit 2 to the required value, and vary that of circuit 3 until no beats are heard between the heterodyne frequency and the constant audio-frequency. (ii) Close circuit 1 and vary C_1 until the beats again disappear, i.e. the frequency of circuit 2 is again at its original value. It is shown in Appendix 2 that this gives the resonant value of C_1, i.e.

$$\omega L_1 = 1/(\omega C_1)$$

(iii) Displace C_1 by a small amount. Beats are heard, due to the consequent change in ω. Synchronism between the heterodyne frequency and the constant audio-frequency is restored by changing C_2. If the process be repeated it will be found that the variation of C_2 required to balance the effect of the variation of C_1 is of the same form as the variation of the frequency in the Willans method, i.e. it will be as shown in Fig. 2, the ordinates being values of C_2 in this case. Thus, for a given setting of C_2 (e.g. that corresponding to the line P_1P_2 in Fig. 2) there will be two values of C_1 (corresponding to the points P_1 and P_2) at which the original frequency $\omega/(2\pi)$ is restored and the beats disappear. These two points coalesce at P, the peak of the variation. If C_2 is increased beyond this point the beats can be reduced to a minimum but not silenced. Thus the critical values C_1', C_1 and C_1'' can be determined with good sensitivity, and R_1 calculated by the same formula as before (which is now exact and not an approximation). Actually,

$$1/C_1' - 1/C_1 = 1/C_1 - 1/C_1''$$

so that it is not really necessary to determine both C_1' and C_1'', though it is a little more accurate to do so.

An alternative formula can be used if desired. Let C_{1a} and C_{1b} be the values of C_1 corresponding to the points P_1 and P_2. Let L_1 be the inductance of the coil, and let

$$\omega L_1 - 1/(\omega C_{1a}) = X_{1a} \ldots, \text{ etc., etc.}$$

Then it is shown in Appendix II that

$$X_{1a}X_{1b} = R_1^2$$

Further, since $\omega L_1 = 1/(\omega C_1)$ as shown above, this formula can be alternatively expressed in the form

$$\frac{1}{\omega^2}\left(\frac{1}{C_1} - \frac{1}{C_{1a}}\right)\left(\frac{1}{C_1} - \frac{1}{C_{1b}}\right) = R_1^2$$

This is possibly a slightly more accurate procedure than that already described, as the points P_1 and P_2 can be determined more precisely than P. On the other hand, this formula will be more subject to the possible error (see Appendix III) due to the variation of the frequency of the first oscillator with resistance.

The variation of C_2 is not necessarily restricted to an amount which would correspond to an audible-range variation of ω, so that a closer coupling between the coils can be used. Further, the sensitivity can be greatly increased if required by beating circuit 3 against a harmonic of circuit 2. The third harmonic is sometimes convenient for this purpose. This use of a harmonic has the further advantage of avoiding any possibility of appreciable interaction between circuits 1 and 3.

Ambiguity in the heterodyne beat is easily avoided, but if a check is required it is sufficient to make sure that variations of C_2 change the pitch in a consistent sense.

The analysis of the method and the derivation of the formula are given in Appendix II.

The simple formula given above is based on the assumption that the frequency of circuit 2 is not appreciably affected by a small change in the effective resistance of coil 2. Actually the frequency will depend to

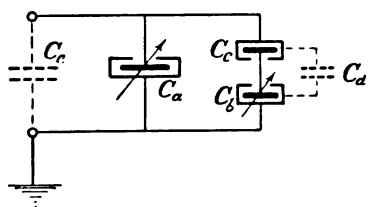

Fig. 3.

some slight extent on the effective resistance of coil 2, and the increase in resistance due to the coupling to the measurement circuit will, therefore, introduce a slight error not taken into account in the simple formula. A fuller discussion of this is given in Appendix III, and it is referred to again in Section (7). The extension of the method to the measurement of impedance in general is described in Appendix IV.

The above is to be regarded as a simplified general account of the method. Practical working details will now be given.

(5) Practical Details.

The actual realization of the method will permit of considerable variation in detail. The following is merely a description of a scheme which has been found to work quite well, but is no doubt open to improvement in detail. The various components will be separately described.

(*a*) *The Measurement Circuit.*—This consists of a convenient wooden stand with sockets and terminals for connection to the coil under investigation, and a system of rigid leads for connecting the measurement condensers. The coil is screened from the measurement condensers by an upright wooden partition which carries a screen of vertical wires. This refinement is made necessary by the high sensitivity of the capacitance adjustment. A mercury link is included in the leads from the coil-mounting to the condensers.

The condenser system is connected as shown in Fig. 3.

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Condenser C_a is a variable condenser with air dielectric, quartz insulation and a maximum capacitance of about 600 $\mu\mu$F, while C_b is a similar condenser with a maximum capacitance of about 200 $\mu\mu$F. C_c is one of a series of small fixed condensers built up of aluminium plates about 1¼ in. square contained in a screening metal case connected to one electrode, their capacitances being of the

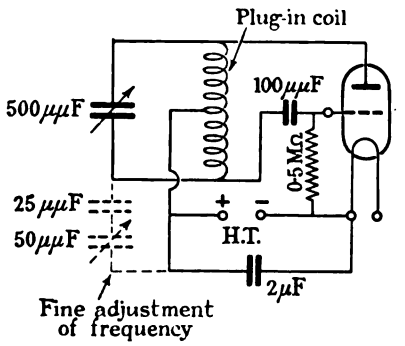

FIG. 4.

order of 10, 25, 50, 150 and 200 $\mu\mu$F. The arrangement may appear complicated, but it must be realized that in some cases the capacitance change which has to be measured may be as small as 1 $\mu\mu$F. With $C_c = 10$ $\mu\mu$F this will mean a change of 10 to 20 scale divisions on C_b, so that even this exceedingly small change can be read and repeated to fairly good accuracy.

Where a large number of measurements are contemplated it is worth while to plot curves of $C_cC_b/(C_c + C_b)$ against C_b for the set of fixed values of C_c.

Body-capacitance effects must, of course, be avoided. The screen already mentioned is for this purpose, and the use of a long handle for C_b will eliminate any residual effect.

(*b*) *The First Oscillator.*—The chief requirement to be satisfied by this oscillator is that the frequency shall be stable and as far as possible independent of the resistance of the oscillating circuit. In addition, a very fine adjustment of the capacitance is required. The series-fed Hartley circuit has been found satisfactory and has been used up to the present. The circuit diagram is shown in Fig. 4. A full account of this type of oscillator has already been published*; the modulator attachment described therein is not required in the present application. A convenient fine adjustment of tuning capacitance is obtained as already described in connection with the measurement circuit, i.e. by means of a small fixed and a relatively large variable capacitance in series.

It seems likely that there is room for improvement in this element of the assembly, and some further work will be carried out on the subject. It was decided not to build a special oscillator for the purpose until the best type of circuit had been determined. When one is made the fine capacitance adjustment will probably be incorporated in it. In the meantime the oscillator described is simple in type and answers the purpose reasonably well, though the variation of frequency with coil resis-

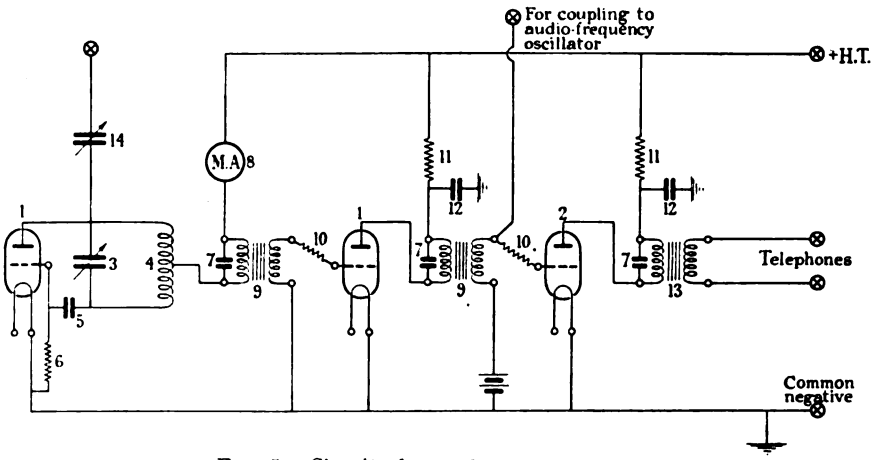

FIG. 5.—Circuit of second or control oscillator.

1.—High-frequency valves.
2.—Power valve.
3.—1 000-μμF variable condenser.
4.—Centre-tapped plug-in coil.
5.—100-μμF fixed condenser.
6.—0·25-MΩ resistance.
7.—1 000-μμF fixed condenser.
8.—Milliammeter.
9.—Audio-frequency transformers.
10.—0·5-MΩ resistance.
11.—2 000-ohm resistance.
12.—2-μF condenser.
13.—Telephone transformer.
14.—20-μμF variable condenser.

C_d and C_e (shown dotted) represent the capacitance of the leads and the mutual capacitance between the cases of C_b and C_c. These must, of course, be determined, and this is easily done by measuring the resonant capacitance of a coil with a number of modifications of the connections of the condenser system. It is not considered necessary to describe these simple measurements in detail. The total capacitance is obviously

$$C_a + C_d + C_e + C_cC_b/(C_c + C_b)$$

tance begins to be appreciable at frequencies lower than 10^5 cycles. The significance of this is discussed more fully in Appendix III.

(*c*) *Second or Control Oscillator.*—Any simple oscillator of stable frequency with provision for headphones in the anode circuit can be used, but a somewhat more elaborate arrangement has actually been constructed. The circuit diagram is given in Fig. 5. The set is com-

* *Experimental Wireless*, 1927, vol. 4, p. 722.

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pletely enclosed in a metal case, except for the coil, which is plugged into sockets on top of the case. It will be seen that the set embodies a single stage of low-frequency amplification. This enables sufficient headphone intensity to be obtained with an exceedingly loose coupling to the first oscillator, a very desirable condition. The retroaction condenser is made variable, because loose coupling between the coil and the driving valve reduces frequency variation. (In the first oscillator the retroaction condenser is made fixed and fairly large, as a loose coupling in this case would tend to increase the frequency-coil resistance variation.) A milliammeter is included in the anode circuit of the first valve to indicate the oscillating condition, and the retroaction condenser is adjusted to the minimum value consistent with oscillation.

Provision is made for capacitance coupling between the second and first oscillators should this for any reason be more convenient (see condenser 14 in Fig. 5). The use of an output transformer for the headphones permits of the earthing of the telephone leads. This reduces undesired stray capacitance effects.

(*d*) *Audio-frequency Oscillator.*—Any of the usual arrangements will suffice for this, provided the frequency

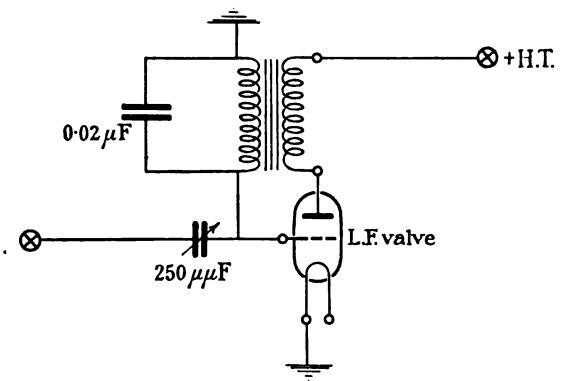

FIG. 6.—Circuit of audio-frequency oscillator.
Grid winding:—1 000 turns No. 36 S.W.G. (about 5 cm diameter).
Anode winding:—100 turns No. 22 S.W.G. (about 5 cm diameter).

is reasonably stable. The oscillator actually used at present is contained in a small metal box about 10 in. cube. The circuit is as shown in Fig. 6. The actual frequency is largely a matter of choice; about 500 cycles is quite suitable. Provision is made for changing the frequency, if required, by means of small fixed condensers on a plug-in mounting.

The coupling between the audio-frequency oscillator and the second oscillator is by capacitance between one terminal of the audio-frequency coil and the grid of the output valve. A variable condenser (250 $\mu\mu$F) is incorporated in the audio-frequency oscillator for this purpose.

In a first design the audio-frequency oscillator was incorporated in the second radio-frequency oscillator, but it was found that stray couplings gave far too high a minimum audio-frequency intensity. The present arrangement was adopted in order to maintain complete control of the audio-frequency intensity. This should in every case be comparable with the beat-frequency intensity.

(6) DETAILS OF THE MEASUREMENT.

The following are practical points of detail in the measurement:—

(i) To ensure stability of frequency in the first and second oscillators it is advisable that these should be switched on some hours before a measurement is made.

(ii) Capacitance couplings between the first and second oscillators will not produce any appreciable error in the resistance measurements, provided that the coupling capacitances are small compared with the tuning capacitances. To make such unavoidable capacitance couplings quite definite in location it is desirable that constant-

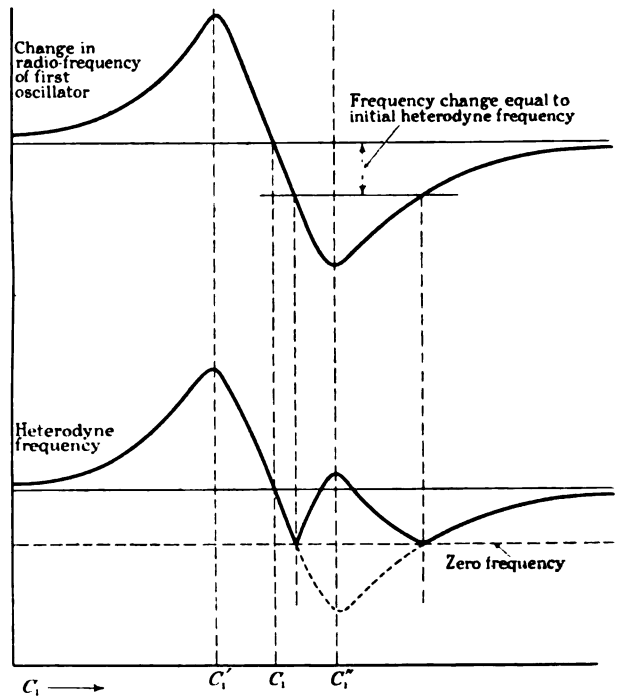

FIG. 7.

potential points should be earthed to the cases of the condensers and the screening cases of the oscillators. It was found desirable to use screened flexible cord with the covering earthed for the coupling between the audio-frequency oscillator and the second radio-frequency oscillator.

(iii) The coupling between the measurement circuit and the first oscillator should be as close as is consistent with the regular behaviour of the latter. If the coupling is too close, tuning of the measurement circuit through resonance may absorb too much power from the oscillator, causing a discontinuous variation in the beat frequency. A suitable coupling is readily found by trial. It will be found that the closer the coupling the larger will be the frequency variation. For a given coupling the frequency variation will increase as the coil resistance decreases. All preliminary adjustments are best made with the audio-frequency oscillator disconnected from the second oscillator. A switch or link can be provided in the coupling lead for this purpose. It is better not to switch

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out the audio-frequency oscillator, as this impairs its frequency stability. As the measurement circuit is tuned the actual frequency variation may extend beyond the audible limits, particularly if the second oscillator is arranged to beat with a harmonic of the first for the sake of sensitivity. As already pointed out, however, this does not matter with the present method of determining the critical capacitances, but it should be observed that with a large frequency-change the variation of the audible beat-frequency in the preliminary adjustment will not necessarily follow exactly the form of the variation of ω. The radio-frequency change will be as in the top curve of Fig. 7. If the total decrease in radio frequency exceeds the initial beat-frequency, the beat frequency will pass through zero at the two settings of C_1 where the decrease in the radio frequency is equal to the initial beat-frequency, and will rise to a new maximum (instead of falling to a minimum) between these points, as shown in the lower curve of Fig. 7. The effect is quite easy to recognize and need cause no confusion, though it will call for a little more care in determining the centre setting.

A suitable coupling having been found, the measurement circuit is opened and the first oscillator is adjusted to give zero beats between the heterodyne frequency and the fixed audible frequency, the heterodyne frequency rising with increase of first-oscillator tuning-capacitance. The measurement circuit is closed and the setting of the measurement-circuit capacitance for zero beats is found by coarse adjustment of the main tuning condenser C_a and fine adjustment on the auxiliary variable condenser C_b (Fig. 3). A rough trial should now be made to ensure that the fixed auxiliary condenser C_c is of suitable magnitude. The smallest value should be taken consistent with finding both critical capacitance values on the scale of C_b. This having been done, C_b is decreased until the highest heterodyne-frequency point is reached and then the first-oscillator tuning-capacitance is reduced to restore the note to its original pitch, i.e. to restore the zero-beat condition. If this does not correspond to the peak of the capacitance variation curve there will be some other zero-beat point for a small displacement of C_b (cf. the points P_1 and P_2 in Fig. 2). By successive changes of the oscillator tuning capacitance and of C_c, the peak value (C_1'' of Fig. 2) is easily determined. If the oscillator tuning capacitance be further increased beyond this point, adjustment of C_b will give a point of minimum beat frequency but no zero-beat points. The second critical point C_1' is found in a precisely similar manner, but in this case the heterodyne frequency falls instead of rising. (The calculation of C_1' and C_1'' from the observed readings of C_b has already been discussed above.)

For the calculation by means of the formula

$$R = \frac{1}{2\omega}\left(\frac{1}{C_1'} - \frac{1}{C_1''}\right)$$

it is necessary to know ω. If the inductance and self-capacitance of the measurement coil (L_1 and C_0) are accurately known, ω can be calculated from the formula

$$\omega^2 L_1(C_1 + C_0) = 1$$

where C_1 is the centre zero-beat-point capacitance (see Fig. 2). If they are not known a wavemeter will be required, and a very sensitive determination of the setting of the wavemeter condenser is obtained by coupling loosely to the first oscillator and finding the centre zero-beat-point exactly as with the measurement circuit. The latter must, of course, be open-circuited during this proceeding.

The alternative method already referred to in Section (4) can be used if desired, namely the determination of the two zero-beat points P_1 and P_2 (Fig. 2) for some given value of C_2 below the peak value. If C_{1a} and C_{1b} be the values of C_1 for any such pair of points, and C_1 be the centre zero-beat capacitance, then, as shown in Appendix II,

$$\frac{1}{\omega^2}\left(\frac{1}{C_1} - \frac{1}{C_{1a}}\right)\left(\frac{1}{C_1} - \frac{1}{C_{1b}}\right) = R_1^2$$

Whichever formula is used, a correction for the frequency-resistance variation of the first oscillator can be applied if this is appreciable in relation to the desired accuracy of measurement. This correction is considered in more detail in Appendix III.

The above somewhat lengthy description may give the impression that the method is complicated and difficult. With a very little practice, however, it will be found both quick and comparatively simple. Only four readings have to be taken, including the measurement of the wavelength, and if the capacitance curves already mentioned be drawn the calculation is a matter of minutes only.

(7) Frequency-Range and Accuracy Obtainable.

The accuracy obtainable cannot be specified as an overall percentage figure, as it will depend on the accuracy with which the variable condensers can be read and on the total readings of these condensers. It has been found that the capacitance changes can be repeated under different conditions with a maximum variation of 0·05 $\mu\mu$F, and the total change will vary from about 1·5 $\mu\mu$F for very low-resistance coils to about 10–20 $\mu\mu$F for coils of higher resistance. This corresponds to a variation in accuracy of from 0·25 per cent to about 3 per cent. This is the principal factor in question, but there is room for improvement here and this can be obtained, if required, by modifying the condenser system.

The variation in the frequency of the first oscillator with resistance is also a possible source of inaccuracy, as already explained, but where maximum precision is required this error is determinable and can be eliminated. Here, also, there is probably room for improvement by suitable modification of the design of the first oscillator, and work will be continued in this direction.

The representative selection of measurements listed in Table 1 will give some idea of the overall accuracy obtainable. It will be observed that the consistency is generally better than 2 per cent. The absolute accuracy appears to be not quite as good as this, judging by the measurement of inserted resistances of very fine wire (measurements 2, 4 and 6), but it must be remembered that the self-capacitance of the coil and associated leads will partially short-circuit a resistance inserted between

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coil and condenser, leading to a slight reduction of its effective value.* Measurements of high resistances of the Loewe vacuum pattern, connected in parallel with a coil as described in Appendix IV, gave agreement with the measured d.c. value to better than 2 per cent.

It will be observed that some measurements were made at a wavelength as short as 28·4 metres (about 10 million cycles per sec.). Given a satisfactory earthing system the method is quite easily workable at such frequencies, though it requires a little more care in experimental details. It was, for instance, found preferable to use a loud-speaker instead of telephones, to avoid minute changes in the capacitance of the system due to movement of the telephone leads.

There appears to be no lower radio-frequency limit, though with coils of ordinary dimensions the error due to variation of the frequency of the first oscillator with resistance will become more pronounced at the lower end of the radio-frequency range.

Attention is particularly directed to measurements 7 and 8. These are typical of a series of observations made to determine whether the stray-capacitance coupling between the measurement circuit and the first oscillator would cause any error. It was found that elimination of such couplings by screening made no difference to the measurements.

It is not intended to invite controversy by making extravagant claims for the method here described in comparison with existing methods, such as resistance or reactance variation. It may fairly be pointed out, however, that methods involving current or potential measurement are usually liable to errors due to variation of the effective induced e.m.f. caused by reaction on the source, or by modifications of the geometry of the system during the measurement. The present method is free from such errors.

TABLE 1.

No.	Subject of measurement	λ	C_1	C_2	C_3	R	Remarks
		metres	$\mu\mu$F	$\mu\mu$F	$\mu\mu$F	ohms	
1	Coil A.5.	956·4	524·80	527·40	530·09	4·86	
			524·04	526·70	529·35	4·85	
			524·02	526·70	529·34	4·86	
2	Coil A.5. with a fixed resistance in series	956·4	520·60	526·60	532·73	11·10	Therefore inserted resistance = 6·33 (measured d.c. value 6·66).*
			520·85	526·90	532·93	11·22	
			519·74	525·80	532·0	11·26	
3	Coil A.1.	190·3	514·73	516·20	517·51	0·53	Mean = 0·54.
			514·33	515·80	517·29	0·56	
			514·76	516·20	517·63	0·54	
			514·58	516·00	517·44	0·54	
4	Coil A.1. with fixed resistance in series	190·3	496·58	514·40	533·08	6·96	Therefore inserted resistance = 6·42 (measured d.c. value = 6·66).*
5	Coil S.0.	28·4	242·91	244·90	246·84	0·49	
			243·38	245·40	247·38	0·50	
			240·96	242·80	244·80	0·49	
6	Coil S.0. with fixed resistance in series	28·4	223·00	251·30	283·72	7·24	Therefore inserted resistance = 6·76 (measured d.c. value = 6·8).
			223·90	251·90	285·80	7·35	
			221·38	249·10	280·24	7·15	
7	Coil A.5. with a wire screen round 1st oscillator coil	925	474·72	477·16	479·60	5·26	
8	Coil A.5. with no screen round 1st oscillator coil	925	474·95	477·38	479·83	5·26	

* Corrected for the measured self-capacitances of the coils and with a conservative estimate of 2 $\mu\mu$F for the leads and wiring, the measured values become 6·50 for No. 2 and 6·64 for No. 4.

In conclusion, the authors wish to express their indebtedness to Mr. A. C. Gordon-Smith of the National Physical Laboratory, who constructed much of the apparatus used in this investigation and made practically the whole of the measurements and observations involved. The work was carried out as part of the programme of the Radio Research Board, and is published by permission of the Department of Scientific and Industrial Research.

APPENDIX I.

ANALYSIS OF THE WILLANS METHOD.

Referring to Fig. 1, it is assumed that any small change in the effective reactance of circuit 2 will give

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rise to a proportional change in the self-maintained frequency of oscillation, or

$$\delta\omega \propto \delta X_2$$

When circuit 1 is closed

$$\delta X_2 = M^2\omega^2 X_1/(R_1^2 + X_1^2)$$

If it now be assumed that ω is constant and that X_1 alone is varied, the right-hand side reaches critical values when

$$R_1^2 = X_1^2$$

Again, assuming that the small change in ω can be neglected, we have

$$R_1 = \pm\,[\omega L_1 - 1/(\omega C_1)]$$

and if C_1' and C_1'' be the two solutions of this equation

$$-\omega L_1 + \frac{1}{\omega C_1'} = R_1$$

and

$$+\omega L_1 - \frac{1}{\omega C_1''} = R_1$$

Therefore

$$R_1 = \frac{1}{2\omega}\left(\frac{1}{C_1'} - \frac{1}{C_1''}\right)$$

Note, however, that the assumption of constant frequency comes in at two points. The first is in the deduction of the critical condition $R_1^2 = X_1^2$, and here the error introduced will be very small unless the circuit has an exceedingly high L/C ratio. The second point at which the assumption comes in is where C_1' and C_1'' are determined. For the more accurate determination it will be necessary to allow for the frequency-changes $\pm\,\delta\omega$ corresponding to the two capacitance values C_1' and C_1''. This will lead to the more accurate expression

$$R_1 = \frac{1}{2\omega}\left(\frac{1}{C_1'} - \frac{1}{C_1''}\right)\left(1 + \frac{\omega L_1}{R_1}\frac{\delta\omega}{\omega}\right)$$

from which the error due to the omission of the second factor can be estimated. Assuming the values 100 for $\omega L_1/R_1$ and 1/100 for $\delta\omega/\omega$, it will be seen that the error may easily be as much as 100 per cent.

The error is smaller at the higher radio-frequencies and with coils of higher decrement. It would be possible to allow for the error, but this would complicate the method considerably.

APPENDIX II.

Analysis of Present Method, neglecting Resistance-Frequency Correction.

For the present purpose it will be assumed that the frequency of the self-maintained first oscillator is not affected by the resistance of the oscillating circuit, being given by

$$\omega^2 L_2 C_2 = 1 \quad . \quad . \quad . \quad . \quad . \quad (1)$$

i.e.

$$\omega L_2 - 1/(\omega C_2) = X_2 = 0 \quad . \quad . \quad . \quad . \quad (2)$$

The effect of a possible departure from this formula due to resistance will be considered in Appendix III.

The condition for constant frequency when circuit 1 is closed is

$$\omega L_2 - \frac{1}{\omega C_2'} - \frac{M^2\omega^2}{R_1^2 + X_1^2}X_1 = 0 \quad . \quad . \quad (3)$$

where

$$X_1 = \omega L_1 - 1/(\omega C_1) \quad . \quad . \quad . \quad . \quad (4)$$

It is clear that when $X_1 = 0$, $C_2' = C_2$, i.e. the value of C_1 such that ω is not changed by closing circuit 1 is given by

$$X_1 = \omega L_1 - 1/(\omega C_1) = 0 \quad . \quad . \quad . \quad (5)$$

Equation (3) can be written in the form

$$X_1^2 - A'X_1 + R_1^2 = 0 \quad . \quad . \quad . \quad . \quad (6)$$

where

$$\frac{1}{A'} = \frac{1}{M^2\omega^2}\left(\omega L_2 - \frac{1}{\omega C_2'}\right) \quad . \quad . \quad . \quad (7)$$

If X_{1a} and X_{1b} be the two roots of equation (6) for a given value of C_2' (i.e. the value of X_1 corresponding to the points P_1 and P_2, Fig. 2), then the product of these roots is R_1^2, i.e.

$$X_{1a}X_{1b} = R_1^2 \quad . \quad . \quad . \quad . \quad . \quad (8)$$

Applying equation (5), this can also be expressed

$$\frac{1}{\omega^2}\left(\frac{1}{C_1} - \frac{1}{C_{1a}}\right)\left(\frac{1}{C_1} - \frac{1}{C_{1b}}\right) = R_1^2 \quad . \quad . \quad (9)$$

The peak values of X_1 occur when the roots of equation (6) are equal, i.e. when

$$X_1'^2 = R_1^2$$

or

$$X_1' = \omega L_1 - 1/(\omega C_1') = -R_1 \quad . \quad . \quad (10)$$

and

$$X_1'' = \omega L_1 - 1/(\omega C_1'') = +R_1 \quad . \quad . \quad (11)$$

Therefore

$$\frac{1}{2\omega}\left(\frac{1}{C_1'} - \frac{1}{C_1''}\right) = R_1 \quad . \quad . \quad . \quad . \quad (12)$$

Either of the equations (9) or (12) can be used for the determination of R_1, but, for reasons given in Appendix III, the second is the more accurate.

It is of interest to note that if C_2' and C_2'' be the values of C_2 corresponding to C_1' and C_1'', then from equations (2), (3), (10) and (11)

$$\frac{1}{\omega}\left(\frac{1}{C_2'} - \frac{1}{C_2''}\right) = \frac{M^2\omega^2}{R_1} \quad . \quad . \quad . \quad (13)$$

so that if C_2' and C_2'' are measured at the same time as C_1' and C_1'', M can be determined. Also, the left-hand side of equation (13) is the increase in the effective resistance of circuit 2 when circuit 1 is tuned to resonance.

APPENDIX III.

Effect of Resistance on Frequency of Self-Maintained Oscillator.

The effect of the resistance of the oscillating circuit on the self-maintained frequency of oscillation is likely to depend considerably on the type of oscillating circuit, the voltage and other conditions.

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With the series-fed Hartley circuit arrangement at present being used for the first oscillator it was observed that changes of the coil resistance (by inserted resistance) did not materially affect the mean anode current until the resistance was such as nearly to stop the oscillation. Assuming this to imply that the effective resistance of the valve was not materially affected by the coil resistance, it seemed likely that a relationship of the form

$$\omega^2 L_2 C_2 = 1 + aR_2 \quad . \quad . \quad . \quad . \quad (14)$$

might hold, a being a very small quantity compared with 1. This would give, as the relation between changes δC_2 and δR_2 for constant ω,

$$\delta C_2 = aC_2 \delta R_2 \quad . \quad . \quad . \quad . \quad . \quad (15)$$

The relation was investigated experimentally by determining the changes δC_2 corresponding to inserted resistances δR_2, the frequency being maintained constant by means of the audio-frequency beat method used for the resistance measurement. The condenser arrangement shown in Fig. 3 was used to permit of the measurement of the exceedingly small capacitance changes (of

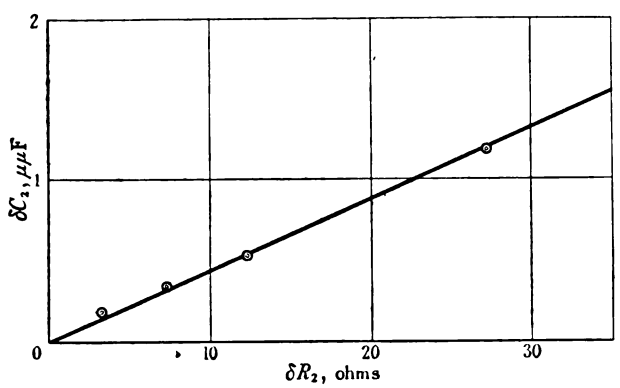

FIG. 8.

the order of 0·1 $\mu\mu$F). A typical result, shown in Fig. 8, is seen to confirm the suggested formula for the frequency.

The effective resistance of circuit 2 is increased by an amount

$$\frac{M^2\omega^2}{R_1^2 + X_1^2} R_1 \quad . \quad . \quad . \quad . \quad . \quad (16)$$

where $X_1 = \omega L_1 - 1/(\omega C_1)$, on account of the coupling to circuit 1. This will produce a change in frequency in accordance with the above formula, i.e.

$$\omega^2 L_2 C_2 = 1 + aR_2 \quad . \quad . \quad . \quad . \quad (17)$$

or

$$\omega L_2 - \frac{1}{\omega C_2} - \frac{aR_2}{\omega C_2} = 0 \quad . \quad . \quad . \quad (18)$$

The analysis must take account of this effect. It may be pointed out, however, that the critical capacitances C_1' and C_1'' will be affected in the same sense and the difference between them will, therefore, be affected to a much smaller degree. This is confirmed by the following analysis.

Using the same symbols as heretofore, and referring to Fig. 1, the initial condition of circuit 2 (circuit 1 open) is given by

$$\omega L_2 - \frac{1}{\omega C_2'} - \frac{aR_2}{\omega C_2'} = 0 \quad . \quad . \quad . \quad (19)$$

When circuit 1 is closed the reactance is decreased by

$$\frac{M^2\omega^2}{R_1^2 + X_1^2} X_1 \quad . \quad . \quad . \quad . \quad . \quad . \quad (20)$$

and the resistance is increased by

$$\frac{M^2\omega^2}{R_1^2 + X_1^2} R_1 \quad . \quad . \quad . \quad . \quad . \quad . \quad (21)$$

Therefore, if C_2' be the new value of C_2 which restores ω to its original value

$$\omega L_2 - \frac{1}{\omega C_2'} - \frac{aR_2}{\omega C_2'} - \frac{M^2\omega^2}{R_1^2 + X_1^2}\left(X_1 + \frac{aR_1}{\omega C_2'}\right) = 0 \quad (22)$$

Comparing this with (19), the centre-point value of X_1 is

$$-aR_1/(\omega C_2') \quad . \quad . \quad . \quad . \quad . \quad (23)$$

Putting $$\frac{1}{A'} = \frac{1}{M^2\omega^2}\left(\omega L_2 - \frac{1}{\omega C_2'} - \frac{aR_2}{\omega C_2'}\right) \quad . \quad . \quad (24)$$

equation (22) can be written

$$X_1^2 - A'X_1 + R_1^2 - A'\frac{aR_1}{\omega C_2'} = 0 \quad . \quad . \quad (25)$$

If X_{1a} and X_{1b} be the values of X_1 corresponding to the points P_1 and P_2 (for a given value of C_2')

$$X_{1a}X_{1b} = R_1^2 - A'\frac{aR_1}{\omega C_2'} \quad . \quad . \quad . \quad (26)$$

For the peak values

$$X_1'^2 = R_1^2 - A'\frac{aR_1}{\omega C_2'} \quad . \quad . \quad . \quad . \quad (27)$$

where C_2 now means the value of C_2 corresponding to C_1'.

Also, by taking the sum of the roots of equation (25) for the critical case

$$2X_1' = A' \quad . \quad . \quad . \quad . \quad . \quad (28)$$

Therefore $$X'^2 = R_1^2 - 2X_1'R_1 a/(\omega C_2') \quad . \quad . \quad (29)$$

and

$$X_1' = -k'R_1 \pm R_1\sqrt{(1 + k'^2)}, \text{ where } k' = a/(\omega C_2') \quad (30)$$

The positive sign corresponds to X_1' and the negative to X_1'', so that

$$X_1'' - X_1' = \frac{1}{\omega}\left(\frac{1}{C_1'} - \frac{1}{C_1''}\right) = -R_1(k'' - k') + R_1\left[\sqrt{(1 + k'^2)} + \sqrt{(1 + k''^2)}\right] \quad . \quad (31)$$

$$= -R_1(k'' - k') + 2R_1\left[1 + (k'^2 + k''^2)/4\right] \quad (32)$$

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Since k' and k'' are nearly equal small quantities, the first term is negligible. Also

$$(k'^2 + k''^2)/4 = k^2/2, \text{ where } k = a/(\omega C_2) \quad (33)$$

very approximately. Therefore, finally

$$\frac{1}{2\omega}\left(\frac{1}{C_1'} - \frac{1}{C_1''}\right) = R_1\left(1 + \frac{k^2}{2}\right) \quad . \quad . \quad (34)$$

As pointed out above, the main correction term kR_1 disappears in the difference between $1/C_1'$ and $1/C_1''$, and only a residual and much smaller correction term $(1 + k^2/2)$ remains.

For the determination of k it is necessary to know a. This can be measured as explained in the first part of this Appendix. Two typical cases are given:—

(*a*) $a = 8{\cdot}76 \times 10^{-5}$
$C_2 = 480 \times 10^{-12}$ farad,
$\lambda = 2\,140$ metres,
$\omega = 8{\cdot}8 \times 10^5$

Therefore $k = 0{\cdot}207$

and $1 + k^2/2 = 1{\cdot}022$

In this case, therefore, the correction is slightly over 2 per cent.

(*b*) $a = 5{\cdot}7 \times 10^{-5}$
$C_2 = 482 \times 10^{-12}$ farad,
$\lambda = 1\,160$ metres,
$\omega = 16{\cdot}25 \times 10^5$

Therefore $k = 7{\cdot}3 \times 10^{-2}$

and $1 + k^2/2 = 1{\cdot}0053$

i.e. a correction of about $\frac{1}{2}$ per cent.

In general, it may be said that the correction will be smaller the higher the frequency and the larger the capacitance across the coil. (Note that it does not depend on the coupling.) It may be quite appreciable in some cases, but, when high accuracy is required, it is determinable as shown and can be allowed for.

It is possible that the correction factor could be reduced considerably by improvement of the first oscillator in respect of frequency-resistance variation.

APPENDIX IV.

Measurement of Impedance by Present Method.

Briefly, the measurement of impedance by the present method is carried out as follows. Some suitable coil which tunes at the frequency required with a capacitance of a few hundred $\mu\mu$F is mounted on the stand, and the centre or resonant capacitance C_1 and resistance R_1 are measured. The impedance to be measured is connected in parallel with the coil, and the change in C_1 and the change in R_1 are measured. The shunt impedance is then determinable from these readings and a knowledge of the frequency.

The theoretical basis is as follows:—Referring to Fig. 9, the unknown impedance will be represented as a shunt conductance G_s in parallel with a shunt susceptance B_s (which may be positive or negative). The effective resultant impedance to the left of the dotted line in Fig. 9 will be represented by $(R_e + jX_e)$. It should be noted that R_e and X_e will depend on the value of C_1.

Let C_1, C_1', and C_1'' be the centre and peak condenser-settings when the shunt impedance is not connected. Then, as already shown,

$$\omega L_1 - 1/(\omega C_1) = 0$$

and

$$\frac{1}{2\omega}\left(\frac{1}{C_1'} - \frac{1}{C_1''}\right) = R_1$$

Let C_2, C_2' and C_2'' be the centre and peak condenser-settings when the shunt impedance is connected, and let these correspond to resultant values R_e, R_e', R_e''; X_e,

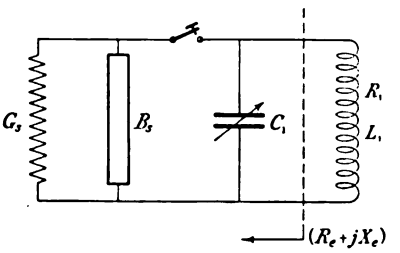

FIG. 9.

X_e', X_e'' of the equivalent impedance elements R_e and X_e. Then, as in Appendix II:—

$$\omega L_1 + X_e = 0$$

and

$$X_e'' - X_e' = 2R_1 + R_e' + R''$$

Also, for any given value of C, since

$$R_e + jX_e = 1/[G_s + j(\omega C + B_s)]$$

Therefore

$$R_e = \frac{G_s}{G_s^2 + (\omega C + B_s)^2}$$

$$X_e = -\frac{(\omega C + B_s)}{G_s^2 + (\omega C + B_s)^2}$$

Thus the full equations for the determination of G_s and B_s are

$$\frac{(\omega C_2 + B_s)}{G_s^2 + (\omega C_2 + B_s)^2} = \frac{1}{\omega C_1}$$

and

$$\frac{(\omega C_2 + B_s)}{G_s^2 + (\omega C_2' + B_s)^2} - \frac{(\omega C_2'' + B_s)}{G_s^2 + (\omega C_2'' + B_s)^2}$$

$$= 2R_1 + G_s\left\{\frac{1}{G_s^2 + (\omega C_2' + B_s)^2} + \frac{1}{G_s^2 + (\omega C_2'' + B_s)^2}\right\}$$

These are impracticably cumbersome, and admit of no simplification without approximation. The following approximations can, however, be made in nearly all practical cases:—

(*a*) G_s^2 is negligible compared with $(\omega C + B_s)^2$ for the values C_2, C_2' and C_2'' of C.

(*b*) $(C_2' \sim C_2'')^2$ is negligible compared with C_2^2.

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The formulæ then become:—

$$\frac{1}{(\omega C_2' + B_s)} - \frac{1}{(\omega C_2'' + B_s)} = 2R_1 + 2\frac{G_s}{(\omega C_2 + B_s)^2}$$

and $1/(\omega C_1) = 1/(\omega C_2 + B_s)$ or $B_s = \omega(C_2 - C_1)$

i.e. $$(C_2'' - C_2')/(\omega C_1^2) = 2R_1 + 2G_s/(\omega C_1)^2$$

Thus if δC be the change in the centre capacitance when the shunt circuit is connected, and if r be the apparent change in the coil resistance

$$B_s = \omega \delta C$$

and

$$G_s = \omega^2 C_1^2 r$$

Assuming these values for B_s and G_s, the condition that G_s^2 is negligible compared with $(\omega C_2 + B_s)^2$ becomes the condition that $\omega^2 C_1^2 r^2$ is negligible compared with 1. Thus the nature of the approximation involved in the above simplified formulæ is very easily determinable for any given measurement.

If the shunt circuit actually consists of, or is to be expressed as, a reactance X in series with a resistance R, then it is easy to show that

$$R = \frac{r}{\omega^2 C_1^2 r^2 + (\delta C/C_1)^2}$$

and

$$-X = \left\{\frac{1}{\omega^2 C_1^2 r^2 + (\delta C/C_1)^2}\right\}\frac{\delta C}{\omega C_1^2}$$

There is, of course, a lower limit to G_s [or an upper limit to $(R_1 + r)$] for this method of measurement to be applicable. If the added resistance due to the shunt circuit is too large, it will not be possible to satisfy the peak conditions with reasonable capacitance changes. In such cases a known capacitance can be interpolated between the shunt circuit and the coil, the capacitance being such as to reduce the total resistance $(R_1 + r)$ to a suitable value for measurement. The resultant shunt impedance is then best calculated in the $(R + jX)$ form. If X' be the reactance component so determined, then the original value X (before the series capacitance was inserted) is obviously $[(X' + 1/(\omega C_0)]$, C_0 being the inserted capacitance. The shunt components G_s and B_s can then be calculated from the formulæ

$$G_s = R/(R^2 + X^2)$$

$$B_s = -X/(R^2 + X^2)$$

Note that this method is applicable to the determination of valve input impedances having a negative resistance component. The effective value of the negative shunt resistance can be reduced by the interpolation of a known capacitance to a value which is well inside the limits of stability of the system, and its actual value can be determined as above from the measured decrease in the apparent resistance of the input coil.

Practical Details.—On general grounds it would seem desirable that the shunt circuit should be removed some distance from the coil and from the first oscillator, to minimize the effect of additional coupling when the shunt circuit is connected. This means fairly long leads —2 metres or so—for the connection of the shunt circuit. A special screened lead was constructed for the purpose, consisting of a copper wire held axially by means of ebonite supports inside a brass tube of about 2 cm diameter. The capacitance of this lead can be determined by a preliminary measurement (it was actually about 30 $\mu\mu$F) and can be allowed for in all subsequent measurements. The procedure will then be to measure the effective resistance and resonant capacitance of the coil with the screened lead connected, connect the shunt circuit and repeat the above measurements. As a check on the method, various measurements were made on shunt circuits built up of known high resistances and known capacitances. For example, a Loewe vacuum resistance of measured d.c. value $0{\cdot}986 \times 10^5$ ohms was connected in parallel with a small fixed air-dielectric condenser of 151 $\mu\mu$F. This system, when measured at a wavelength of about 1 000 metres, gave the values $1{\cdot}00 \times 10^5$ ohms and 151 $\mu\mu$F as the values of the shunt resistance and capacitance respectively. Measurements at higher frequencies gave similar results.

[NOTE.—Since the above paper was written the apparatus has been modified in certain particulars. It was found by measurement that dynatron oscillators, making use of the negative resistance characteristics of screen-grid valves, were superior to the triode oscillator, originally used, in the matter of constant frequency with respect to variation of oscillatory circuit resistance. Such oscillators have therefore been substituted for the triode oscillators for use at frequencies below 10^7 cycles per second. Unfortunately, the dynatron circuit does not oscillate readily at frequencies higher than this.

An output stage has been added to the audio-frequency oscillator (which is also of the dynatron type) with a potentiometer across the secondary of the output transformer. This potentiometer is connected through the telephones in series with the secondary of the output transformer of the second oscillator (Fig. 5). This arrangement makes the audio-frequency independent of the coupling.

The condenser system shown in Fig. 3, though simple and elastic in range, is not entirely satisfactory at frequencies higher than 10 million, as the inductance of the leads cannot be made sufficiently small. It has therefore been replaced by a main condenser with one or other of a set of fine-adjustment condensers connected in parallel with it, with a low-inductance system of connections.]

[The discussion on this paper will be found on page 70.]

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A VARIABLE-CAPACITANCE CYLINDRICAL CONDENSER FOR PRECISION MEASUREMENTS, AND A WAVEMETER FOR SHORT WAVELENGTHS.*

By E. B. MOULLIN, M.A., Associate Member.

(*Paper first received* 18*th October, and in final form* 13*th November*, 1930; *read before the* WIRELESS SECTION 7*th January*, 1931.)

SUMMARY.

The paper describes a form of cylindrical condenser which has been developed for precision measurements at very high radio frequencies. One cylinder is always totally enclosed by the other and the capacitance is varied by altering the distance between the axes of the cylinders. This distance is varied by rotating the inner cylinder about an axle which is eccentric with both cylinders. With the dimensions chosen, the capacitance is increased by 1 $\mu\mu$F by rotating the pointer through 45° from the position of minimum capacitance; the total change of capacitance is 16 $\mu\mu$F. The capacitance values engraved on the scale are determined by calculation and not by experimental calibration. It is shown by electrostatic principles that the end-effect capacitance is likely to be substantially constant as the position of the inner cylinder is varied, and this supposition is supported by several exacting tests by experiment; the validity of the calculated scale is thus vindicated. The condenser can also be used as a very small variable inductance; an important use of this application is for determining the residual inductance of the condenser, which is found to be about 0·05 μH. The paper also includes a tentative design for a short-range, short-wave absorption wavemeter in which the same principles of construction are used.

(1) INTRODUCTION.

An accurately calibrated variable condenser of small capacitance is often required for plotting resonance curves of radio-frequency circuits. At frequencies of the order of 3×10^7 cycles per sec. the total capacitance of the circuit is very small, and the main part of the resonance curve will then be included within a minute change of capacitance. For suppose a simple circuit, of decrement δ, has a total capacitance C_0 at resonance: then ΔC, the change of capacitance necessary to increase the current from half its maximum, up through resonance and down again to the initial value, is given by the formula

$$\delta = \left(\frac{\pi}{2\sqrt{3}}\right)\frac{\Delta C}{C_0}$$

It follows from this that if $\delta = 2$ per cent (power factor = 0·637 per cent) and $C_0 = 35$ $\mu\mu$F, then $\Delta C = 0{\cdot}385$ $\mu\mu$F, and it is therefore necessary to plot resonance curves whose width is only a few tenths of a micromicrofarad, a process which requires a variable condenser having a very open scale. Such a condenser need have neither a large ratio of maximum to minimum capacitance nor an extremely small minimum, but the total change of capacitance must be very small and should at any rate be comparable with the minimum value. In some part of the scale a movement of about 30 angular degrees should correspond to a capacitance change of about 1 $\mu\mu$F. Such an instrument is difficult to calibrate and to do so requires access to a very perfect capacitance bridge with very unequal ratio arms. There is much to be said for a condenser whose change of capacitance can be calculated, so as to obviate the difficulty of experimental calibration.

At very high frequencies the total intentional inductance of the circuit may be of the order of 1 μH, and under such conditions it is extremely important that the residual internal inductance of the condenser should be small and not change appreciably with capacitance.

A parallel-plate condenser does not seem very suitable for the purposes described, whereas a cylindrical condenser has much to recommend it since it is self-screening and has a residual inductance of calculable value.

(2) THE CYLINDRICAL CONDENSER.

It is not desirable to vary the capacitance by inserting one cylinder into the other, because if this is done the inside cylinder is not completely screened and the mechanical motion cannot be readily arranged. The inside cylinder may, however, be enclosed completely and the capacitance varied by moving the axis of the inside cylinder away from the axis of the outside cylinder, thereby making the two cylinders eccentric with one another but with their axes parallel. A cross-section of this arrangement is shown diagrammatically in Fig. 1.

It is well known that the capacitance per unit length of such a system of indefinite length is exactly calculable from the formula

$$C = \frac{1}{2\log_e[\beta + \sqrt{(\beta^2 + 1)}]} \quad . \quad . \quad . \quad (1)$$

where $\beta = (a^2 + b^2 - d^2)/(2ab)$

in which a and b are respectively the radii of the outside and inside cylinders and d is the distance between their axes. The capacitance changes very slowly with d so long as d is a good deal smaller than $(a - b)$, so that there is the possibility of large movement yielding a very small change of capacitance. If the change of capacitance is to be calculable, it is essential that the capacitance due to the ends of the cylinders should not be a function of d. This capacitance cannot be calculated exactly, but a rough approximation to it can be obtained. Consider Fig. 2, which represents the cross-section of an infinitely long flat plate situated within an infinitely long trough; the dimensions AB and BC are supposed to be large compared with g and h. It is well known that the capacitance per unit length of this condenser can be calculated from the formula

$$C = \frac{AB + k_1 g}{4\pi g} + \frac{BC + k_2 h}{4\pi h} + \frac{DC + k_3 g'}{4\pi g'}$$
$$= (1/4\pi)\left[(AB/g) + (BC/h) + (DC/g') + k_1 + k_2 + k_3\right] \quad (2)$$

* Reprinted from *Journal I.E.E.*, 1931, vol. 69 p. 507.

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where k_1, k_2 and k_3 are functions of h/g which can be found by reference to Fig. 3. Suppose that initially $g = g' = 3h$, and that then the plate is gradually moved nearer to one side of the trough until eventually

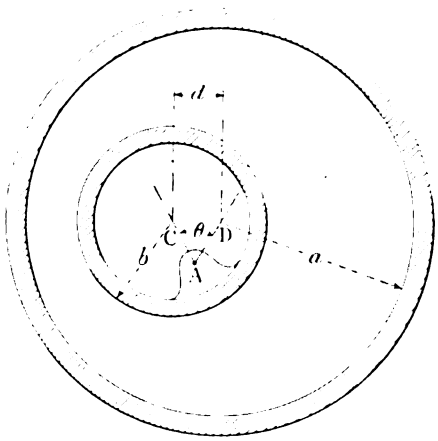

FIG. 1.

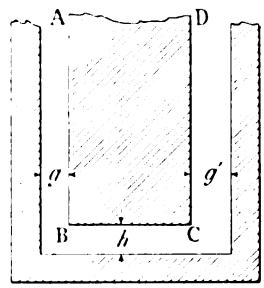

FIG. 2.

$g = \frac{1}{2}h$ and $g' = 5{\cdot}5h$. Fig. 4 shows $(k_1 + k_2 + k_3)$ plotted as a function of g/h; it may be seen that the end capacitance of this condenser remains nearly constant until $g/h = 1$ and that in the whole journey the increase is only 10 per cent. This calculation assumes

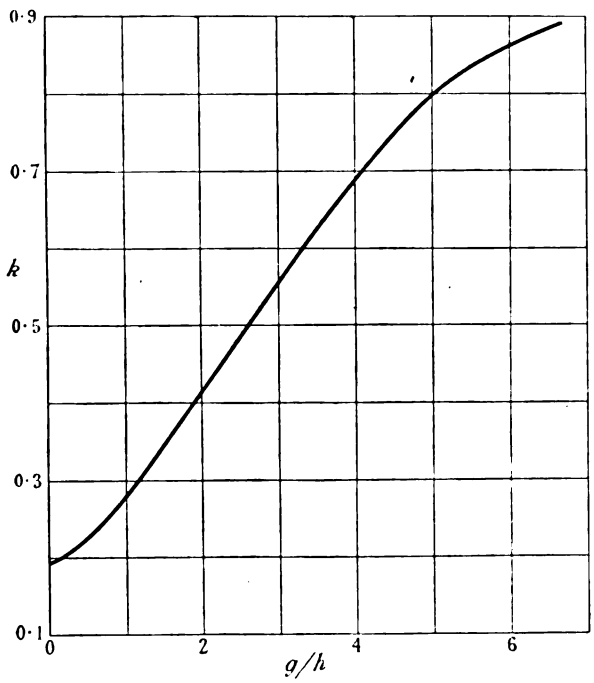

FIG. 3.

that BC is large compared with h, but, as has been shown lately by Mr. Cockcroft,* it need not be very large. Fig. 5, reproduced from Mr. Cockcroft's paper, shows k/h plotted as a function of BC/h for a condenser in which $g = h$, and it may be seen from this that so long as BC is greater than h, k/h is almost independent of BC/h. Corresponding calculations for a cylindrical field are not available, but so long as g and h are smaller than the radius of curvature the calculations for a 2-dimensional field will be substantially correct.

Approximate calculation is still further assisted by a general principle of electrostatics which states that the energy of a given charge is a minimum when the charge has its equilibrium distribution. Except in resonance curves, maxima and minima are never critical, so that the energy corresponding to an approximately correct distribution of charge would differ little from the energy appropriate to the true distribution. If the charge density round the trace of any diametral section of the inner cylinder had the distribution appropriate to the system of Fig. 2, the end capacitance would be substantially independent of the distance between the two axes; the minimum-energy theorem leads us to hope this will be substantially true of the distribution which does in fact obtain round the end of the cylinder. Thus there is reason to expect that the change of capacitance

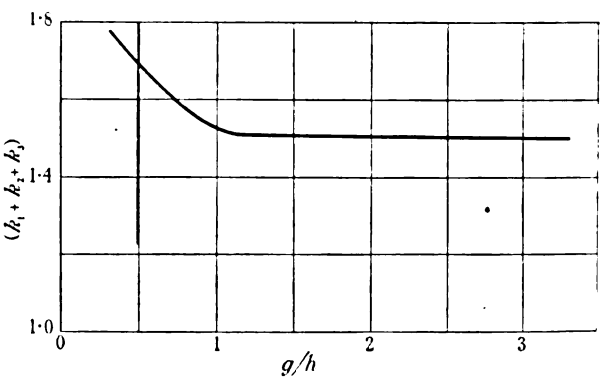

FIG. 4.

due to a change of d (see Fig. 1) may be accurately calculated from formula (1) so long as d is appreciably less than $(a - b)$. When d approaches $(a - b)$, however, the capacitance is increasing at a rapid rate and any change due to the ends will have a small fractional value. Furthermore, in this part of the scale experimental calibration is not difficult.

* *Journal I.E.E.*, 1928, vol. 66, p. 335.

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If end effects are ignored, the product of the capacitance C and inductance L per unit length of two eccentric cylinders is constant and independent of d. The equations of an open-circuited cable of length l may therefore be applied to a long cylindrical condenser, and the input impedance at any frequency can be calculated. Hence, by a well-known formula

$$1/Z = \sqrt{\left(\frac{C}{L}\right)} \tan p\sqrt{(LC)}l$$

$$\doteqdot \sqrt{\left(\frac{C}{L}\right)} p\sqrt{(LC)}l(1 + p^2LCl^2/3), \text{ if } p\sqrt{(LC)}l >> 1$$

$$= pC_s\left[1 + (4\pi^2/3)\,(l^2/\lambda^2)\right] \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (3)$$

Thus the residual inductance will not increase the apparent capacitance by 1·5 per cent until the wavelength is less than 30 times the length of the system. Formula (3) is correct for all values of the eccentricity, because the product LC is constant.

The distribution of charge round the curved surface of the cylinders is known, and the high-frequency resistance of the conductors can therefore be calculated by assuming that the current density penetrates inwards from the surface according to an exponential law. The power factor due to conductor resistance can thus be calculated at any frequency, but the power factor due to dielectric loss in the bushings cannot be calculated. If R_f is the conductor resistance per unit length at any given frequency f, and C is the capacitance, the power factor is

$$P = \tfrac{1}{2}R_f l \,.\, pCl = \tfrac{1}{2}R_f pCl^2$$

and consequently increases as the square of the length of the system. If σ is the current density at some point

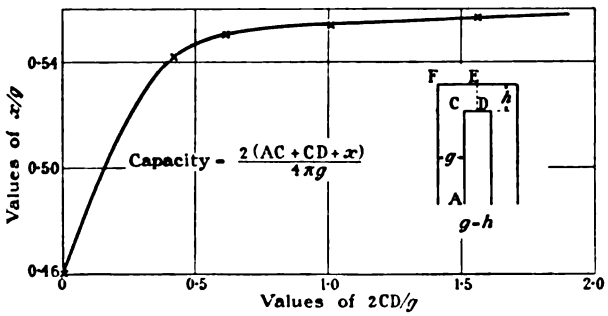

FIG. 5.

on the surface of one cylinder, say the inner, then the resistance loss is proportional to $\int_0^{2\pi} \sigma^2 b\,d\theta$. In this manner it can be shown that the resistance loss is proportional to

$$\frac{1}{2\pi b}\left[\frac{(CB)^2 + b^2}{(CB)^2 - b^2} + x\frac{(DB)^2 + a^2}{(DB)^2 - a^2}\right]$$

where B is the external common inverse point of the two circles forming a cross-section of the condenser, and $x = b/a$. In the instrument to be described, $b/a = \frac{1}{2}$, and the maximum value of the eccentricity makes $d = 0{\cdot}442a$; for these values it may be calculated that $CB = 2{\cdot}35b$ and $DB = 1{\cdot}5a$, whence it is found that the eccentricity increases the resistance by about 30 per cent. It is well known that when the cylinders are concentric

$$R_f = \sqrt{\left(\frac{f\rho}{10^9}\right)} \times \left(\frac{a+b}{ab}\right) \text{ ohms per cm}$$

For copper cylinders in which $a = 2b$, it therefore follows that

$$P = \frac{f^{3/2}\,l^2}{10^9\,b}$$

$$= \frac{1{\cdot}6\,l^2}{10^8\,b} \text{ when } f = 3 \times 10^7 \text{ cycles per sec.}$$

$$= \frac{1{\cdot}6}{10^6} \text{ when } l = 18 \text{ cm and } b = 3 \text{ cm}$$

Thus even at this frequency the power factor due to

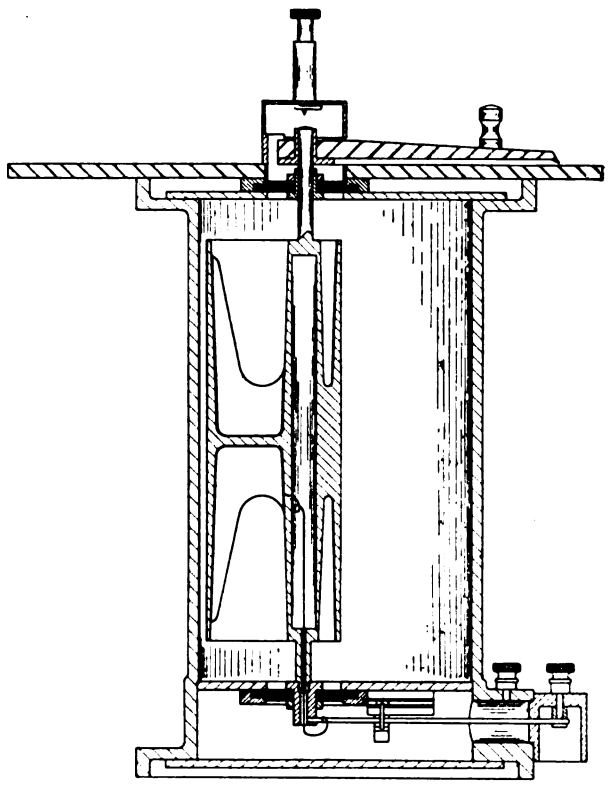

FIG. 6.

conductor resistance is only about 1 per cent of the power factors which have been measured at high frequencies for the best condensers, and power factor due to conductor resistance may therefore be ignored.

Since the cylindrical condenser has a geometrical form for which the capacitance, inductance and resistance can be calculated it has much to recommend it, provided that the mechanical construction can be arranged satisfactorily.

(3) Mechanical Details of the Cylindrical Condenser.

The distance between the axes of the cylinders may be varied conveniently by revolving the inner cylinder about an axis which is eccentric with respect to both

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the inner and the outer cylinder. This arrangement* is shown diagrammatically in Fig. 1, in which the inner cylinder is supposed to revolve about a spindle through the point A. When the inner cylinder is revolved about this axis there is an accompanying variation of the central distance CD, and we have

$$d = \sqrt{[(AC)^2 + (AD)^2 - 2AC \,.\, AD \cos\theta]}$$

This method of altering d lends itself to robust mechanical construction, and it also has the convenience that d changes very slowly with θ when that angle is approaching 180°. It is then that d approaches $(a - b)$ and the capacitance is increasing at a rapid rate.

Fig. 6 shows a cross-sectional drawing of a completed condenser and illustrates the arrangement of the eccentric axle for the inner cylinder. Both cylinders

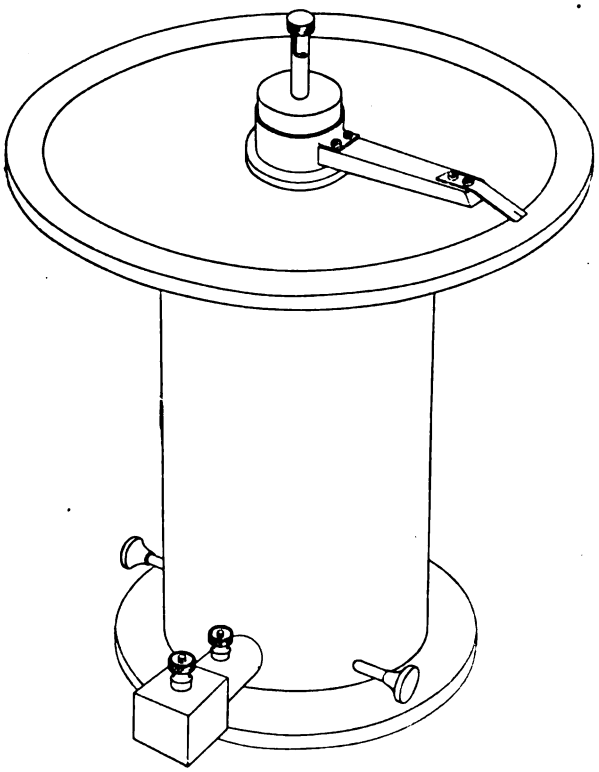

FIG. 7.

are aluminium castings, accurately machined. The axle is carried in brass bearings mounted in mycalex bushes which fit into circular recesses, and these are machined in position so as to ensure that they are on a common axis. The lower bearing is of the footstep type and forms the electrical connection to the rotor. To make this connection more positive than a rubbing contact, a thin copper wire is passed up a central hole in the spindle and soldered to the rotor at one end and to the input terminal at the other. The index pointer is of ebonite and is fixed to a small brass bush which is pinned to the top of the spindle. The top end of the spindle is enclosed by a cylindrical brass cover connected to the outside case. The pointer passes through a slot in this cover and carries a metal ring which forms a door over the slot. The cylindrical cover and the door complete the shielding of the rotor system. The plunger switch mounted on the top of the screen cover (Fig. 6) is for converting the condenser into a variable inductance, and will be referred to later. The footstep bearing is connected to the input terminal by a brass rod passing through the base chamber and finally through a large cylindrical tube formed in the wall of the shell. The base chamber also carries two mica condensers for increasing the range of capacitance. Reference to Fig. 6 will make clear the arrangement of the terminals. A perspective view of the instrument is shown in Fig. 7; the two handles which protrude through the base of the shell are for operating the low-capacitance switches and thus placing the mica condensers in parallel with the system.

* British Patent No. 325802.

(4) Measurements of Capacitance, and Check of Calculated Calibration.

Capacitance measurements were made for a condenser almost identical with that shown in Fig. 6. All linear

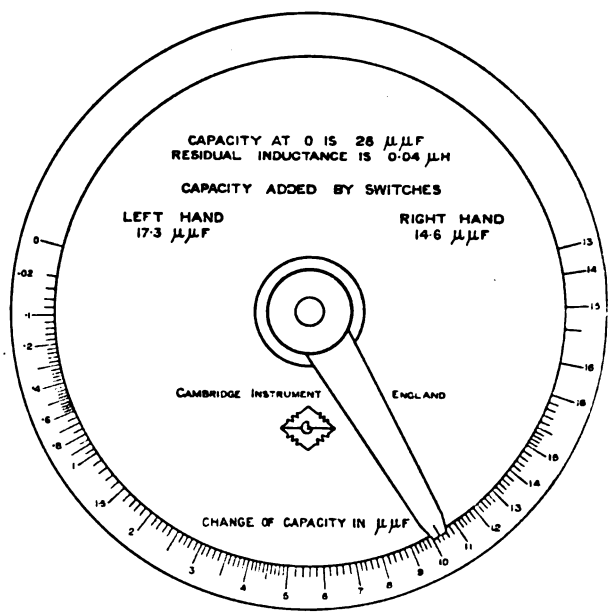

FIG. 8.

measurements were made after assembly, and were taken with the help of a radial pointer and comparator bench capable of being read to 0·01 mm. The length of the rotor was 18 cm, and the minimum capacitance of this length was calculated to be 14·10712 $\mu\mu$F without allowance for end effects. When the cylinders are concentric the side clearance is 3 cm and the end clearance 1 cm. Assuming the fringing to be the same as for the 2-dimensional system of Fig. 2, it may be calculated that the equivalent additional length of the sides is 20 per cent and the equivalent addition to the radius of the top and bottom of the cylinder is 1 per cent. The rotor and bottom bearing having been removed, the capacitance of the terminal system was measured and found to be 4·6 $\mu\mu$F, a value which was

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increased by 2·1 $\mu\mu$F on replacing the bottom bearing. We may therefore make a tentative allowance of 2 $\mu\mu$F for the capacitance of the top bearing and pointer. The values of the capacitances of the components (in $\mu\mu$F) are thus as follows:—

Calculated capacitance of the rotor system ..	14·11
Additional capacitance due to finite sides ..	2·82
Additional capacitance due to top and bottom ..	5·3
Additional capacitance due to terminals ..	4·6
Additional capacitance due to bearings	4·1
Total	30·93

The measured value of the minimum capacitance was 28·2 $\mu\mu$F, so that the corrections have been over-estimated by about 11 per cent.

The corresponding figures ($\mu\mu$F) for a condenser with a smaller rotor are as follows:—

Calculated capacitance of the rotor system ..	10·8
Additional capacitance due to finite sides ..	3·3
Additional capacitance due to top and bottom ..	2·4
Measured capacitance of terminals and bearings..	7·0
Total	23·5

The measured value of the minimum capacitance was 22·0 $\mu\mu$F, so that the corrections have been over-estimated by about 7 per cent.

These figures for the component parts of the minimum capacitance are interesting and they show that the capacitance due to end effects is considerably less than the value estimated from a 2-dimensional problem. These measurements cannot show whether the end capacitance tends to be independent of the position of the rotor.

The change of capacitance with angle was calculated by formula (1). This formula is troublesome for numerical calculation and it involves retaining a great many significant figures. A calculating machine and 7-figure logarithm tables were used to obtain a table of change of capacitance with angle, and the results were engraved on a dial of 10-inch diameter. A reduced facsimile of this dial is reproduced in Fig. 8. The full capacitance change occurs in 180 degrees of arc, and since the calibration must be symmetrical about the diameter through the maximum, or minimum, point the equation relating C and θ must be of the form $C = \theta^2 + \alpha\theta^4 + \beta\theta^6 + \ldots$ For this particular condenser an increase of capacitance 1 $\mu\mu$F is produced by rotating the spindle through 48° 26′ from the position of minimum capacitance. The values of θ were calculated for increases of 0·1, 0·3 and 0·6 $\mu\mu$F and it was found that the equation $C = a\theta^2$ represented the calibration of the first 48 degrees of scale, correct to about one minute of arc. Accordingly, the first micromicrofarad of change was divided into tenths and hundredths according to a square law. The value of θ for a 2-$\mu\mu$F change is one degree less than if the square law held accurately, and the subdivisions of the second micromicrofarad may therefore also be made according to a square law. Between 8 and 14 $\mu\mu$F the calibration is substantially linear and subdivision presents no difficulty, while between 2 and 8 $\mu\mu$F the slight non-uniformity of subdivision can be apportioned by constructing a table of second and third differences.

When the dial has been divided correctly it is essential to set the pointer so that the zero of the scale corresponds precisely to the minimum capacitance. Since the capacitance passes through a maximum at exactly 180 degrees from the position of minimum, it suffices to find the true position of the maximum, and it is for this reason that the dial is scaled for the first 3 micromicrofarads beyond that position. To find the position of maximum capacitance the condenser should be used as the capacitance of a tuned circuit energized from a value generator the frequency of which is set so that the maximum capacitance is just insufficient to bring the circuit to resonance. A current indicator, ammeter or thermionic voltmeter in the circuit will then attain a maximum reading at the position of maximum capacitance. If indicator readings are plotted against capacitance readings a curve will be obtained which will be symmetrical about the ordinate corresponding to maximum capacitance, provided that the pointer is set correctly. If the steep sides of this curve are produced to an apex, this apex should occur exactly half-way between the two highest capacitance values engraved on the scale. Examples of such curves are shown in Fig. 9. The full-line curve is not perfectly symmetrical, which shows that the pointer is not set quite correctly. After shifting the index point slightly in the necessary direction as indicated by the full-line curve, the chain-dotted curve was obtained. The chain-dotted curve is perfectly symmetrical, and the voltmeter reading is the same for a given capacitance as shown by the dia-reading, whether that capacitance is on the ascending or descending side of the maximum. A curve as perfect

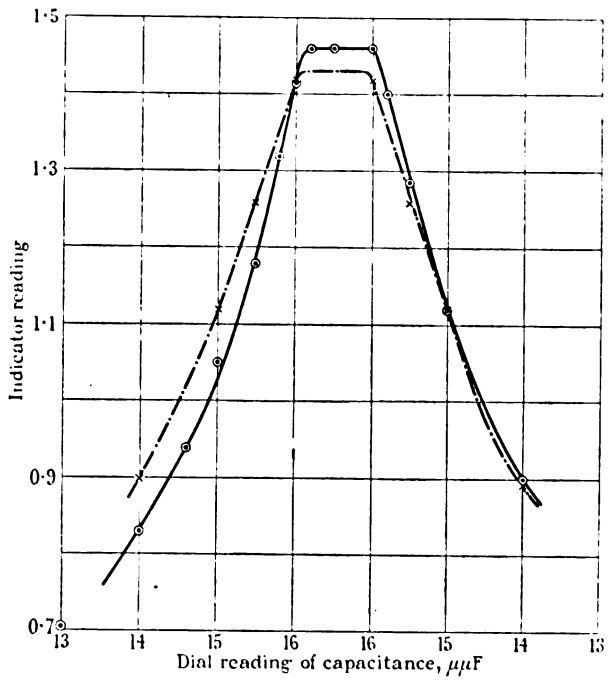

FIG. 9.

Condenser in parallel with 128 $\mu\mu$F and connected to a coil of 500 μH inductance.

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as that shown by the chain-dotted graph in Fig. 9 may not be obtained after one shift of the index point, in which case the process of taking a curve and shifting the pointer must be continued until a perfect curve is obtained. If at any time it is thought desirable to check the zero, this can be done by the process just described.

No capacitance bridge of an accuracy sufficient for testing the calibration experimentally was available, and this has therefore been checked by radio-frequency methods. The tubular condenser was connected in parallel with a uniform-scale variable condenser in which a pointer movement of 3 mm corresponded to a

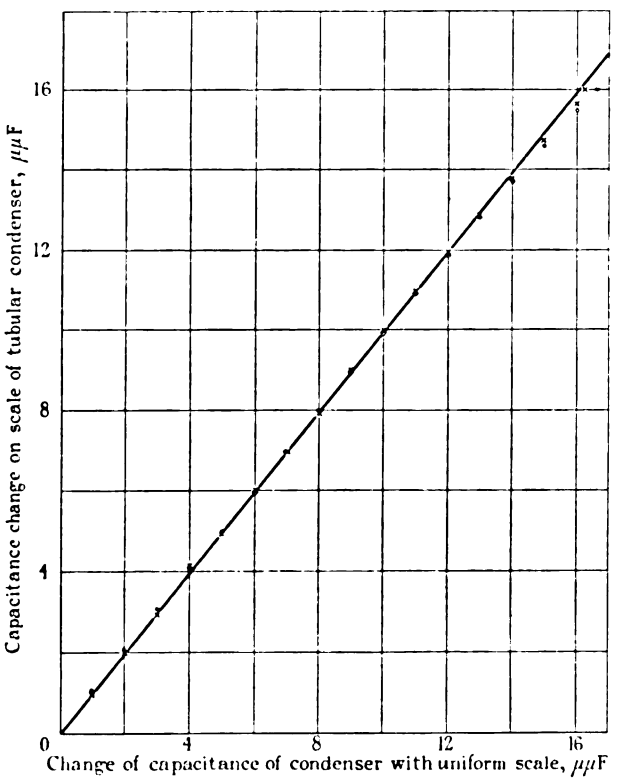

FIG. 10.

capacitance change of 1 $\mu\mu$F. The pointer of the uniform-scale condenser was moved by successive steps of this amount and the decrease of capacitance was replaced by the cylindrical condenser. The test circuit was tuned to a frequency just below resonance with the impressed frequency of 2·45 megacycles per sec. The results of such tests are shown plotted in Fig. 10; the points marked with a cross were obtained from a 16-$\mu\mu$F change in one part of the scale of the uniform condenser and the points marked with a dot were obtained from the same change in another part of the scale. The points do not show more deviation from the line than can readily be accounted for by inaccuracies in setting the pointer of the uniform-scale condenser. At the upper end there is a slight sign of deviation from the line, which could be accounted for in this part of the scale by a very minute inaccuracy in measuring a diameter or an eccentricity. Fig. 10 shows definitely that any regular change of end capacitance which occurs over the whole range of movement is too small to be observed by this method. In the lower part of the scale there is little change of eccentricity, and appreciable change of end capacitance would not be expected. That the points in Fig. 10 lie so near a straight line seems proof that the law of the condenser agrees with that calculated from equation (1), so that there should be no hesitation in accepting the calculated subdivision of the first few micromicrofarads. The absolute value of the total capacitance change can be measured without serious difficulty and this provides a further check on the calculated values.

When a batch of such condensers are made as exact copies of an original, it would seem better to calibrate them by comparison with the original rather than to go through the laborious process of exact linear measurement and re-calculation for each one. Given high-class workmanship each member of the group will be a close copy of the original, and the possible departures of calibration will be small. Electrical methods of comparison would show definitely any systematic differences of calibration, and an intelligent use of difference tables would allow the true calibration to be reproduced from the original with as much accuracy as if it had been obtained *de novo* by linear measurement and calculation from formula (1). This method of procedure would save an enormous amount of labour.

Each of the fixed mica condensers contained in the base chamber is made to have a capacitance of about 14 $\mu\mu$F, or at any rate something less than the maximum change of 16 $\mu\mu$F. Each of these can therefore be calibrated *in situ* by observing the increment of scale reading required to replace it when it is switched out of circuit. In this way the capacitance of each base condenser is consistent with the variable portion, and, moreover, this capacitance can be re-checked at any time, These base-chamber condensers give the instrument a continuous range from 28 to 72 $\mu\mu$F.

(5) Some Applications of the Condenser.

The outstanding use of the condenser in radio-frequency measurements is for plotting resonance curves, and for this purpose it may be used either as the variable condenser of a tuned circuit or as a means of varying the frequency applied to a fixed circuit. It will be required for the first use at high frequencies when the total circuit capacitance is small. One example of this use is shown in Fig. 11, which represents a resonance curve for a circuit having an applied frequency of 7·25 megacycles per sec. and a total capacitance at resonance of 31 $\mu\mu$F. The values of the width of this curve at the heights usually employed for deducing the circuit decrement are 0·86 $\mu\mu$F and 1·54 $\mu\mu$F respectively. Six observation points are shown between 0·8 and 1·0 $\mu\mu$F, and the change of capacitance from one to the next of these is 0·03 $\mu\mu$F. Clearly, a very special condenser is required to permit a resonance curve such as Fig. 11 to be drawn. If ϕ, the lag angle of a circuit, is deduced from the shape of the resonance curve and if $\tan\phi$ is plotted against capacitance change it is well known that the result should be a straight line. This method of rectification has been applied to Fig. 11,

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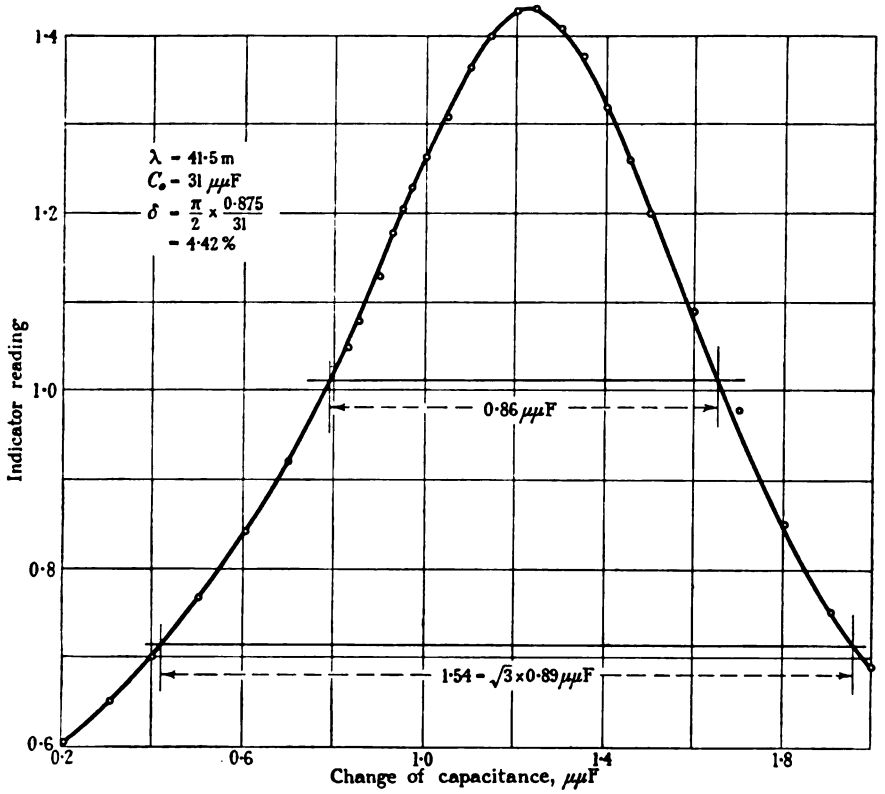

FIG. 11.

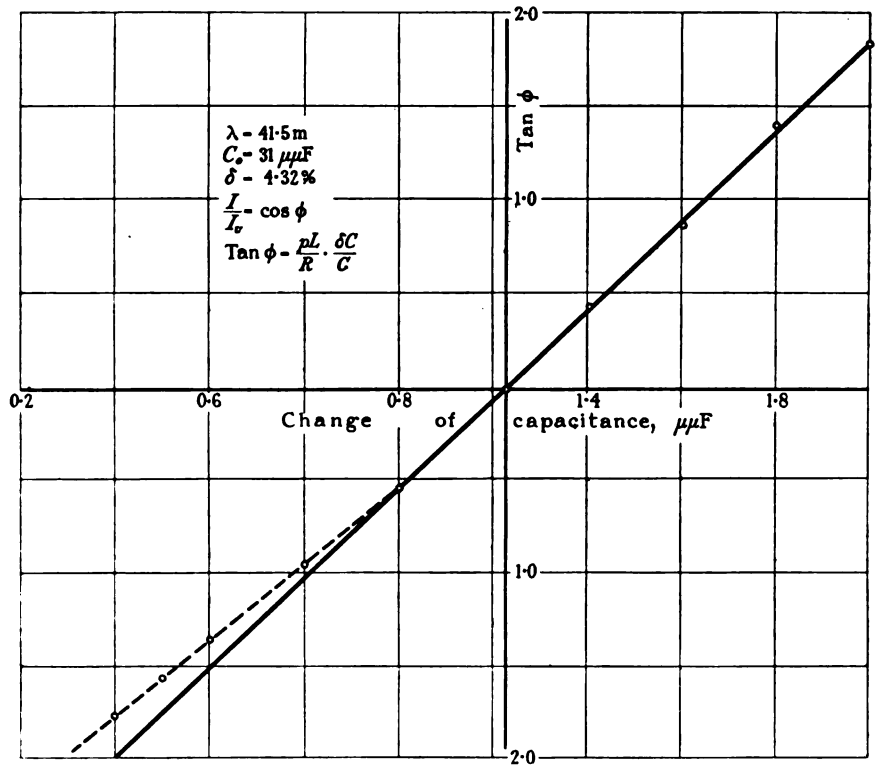

FIG. 12.—Rectified resonance curve.

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and the result is shown in Fig. 12. The points are not exactly on a straight line because the resonance curve has a slight dissymmetry due to the coupling coil to the example is given in Fig. 13, which shows a resonance curve for a circuit having an applied frequency of 26·2 megacycles per sec. ($\lambda = 11{\cdot}4$ m).

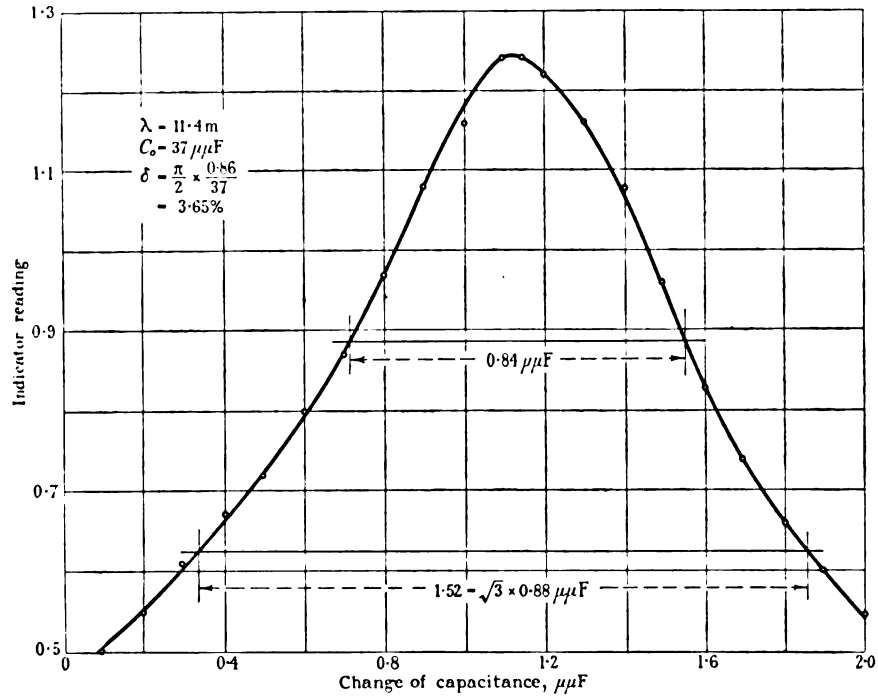

FIG. 13.

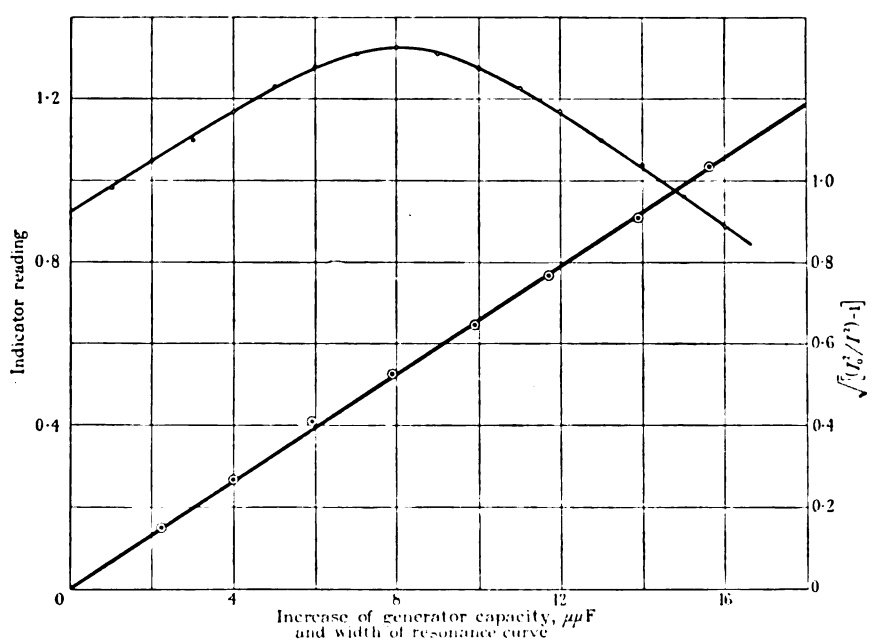

FIG. 14.

λ = 510 m. For receiver circuit, L = 500 μH. Generator capacitance = 570 μμF.

thermionic voltmeter.* Figs. 11 and 12 must surely establish the validity of the scale over the part employed in the tests which they represent. Another similar

* See *Experimental Wireless*, 1930, vol. 7, p. 367.

An example of the second use is shown in Fig. 14, in which the change of capacitance in the generator circuit is plotted against the reading of a thermionic voltmeter coupled to a fixed receiving circuit of natural frequency

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588 kilocycles per sec. It is well known that if Δp is the width across the resonance curve at some fraction $1/x$ of its maximum height, then

$$\Delta p = (R/L)\sqrt{(x^2 - 1)}$$

Assuming that the frequency of the generator circuit is given by the formula $p^2 = 1/(L'C')$,

Then $$2dp/p = dC'/C'$$
$$dC'/C' = (2\delta/\pi)\sqrt{(x^2 - 1)}$$

Thus if the width of the curve shown in Fig. 14 is plotted against $\sqrt{(x^2 - 1)}$, the result should be a straight line through the origin. The straight line in the same figure shows the result of this process. Since the points lie very close to a straight line through the origin the relative values of calibration must be correct over the whole scale. Fig. 15 shows the same process performed for a circuit of natural frequency 1·43 megacycles per sec., and here the capacitance of the generator circuit was increased by the first two micromicrofarads of the scale. The rectified resonance curve is also plotted, and since this is a straight line the relative calibration of the first two micromicrofarads is also vindicated. Thus every test of the scale has shown that the change of capacitance is equal to that calculated from formula (1).

This condenser can also be used in capacitance bridges for measuring the dielectric constant of gases, and in cases where the dielectric constant is only a few parts in a thousand greater than unity. Using an equal-ratio bridge, balance is obtained against an air condenser of capacitance about 100 $\mu\mu$F which is then filled with the gas to be measured, balance being restored by means of the tubular condenser. Given adequate sensitivity in the indicator, the increase of capacitance could readily be measured to 0·01 $\mu\mu$F, while the desirability of an equal-ratio bridge for small-capacitance measurements could still be retained.

(6) Use of the Condenser as an Inductance.

If the spring plunger shown in Figs. 6 and 7 is released the top of the rotor is connected to the outside cylinder, and then the current which enters the insulated terminal passes up the rotor and down the outside cylinder to the case terminal. The system thus forms a self-screening inductance whose value can be varied by revolving the rotor. The calibration in terms of inductance change can be deduced from the capacitance calibration. Thus it follows readily that if the capacitance of the tubular portion is C $\mu\mu$F and the length of the system is l cm, the inductance of the tubular portion when the plunger switch is released is

$$L = (l^2/900C)\ \mu\text{H}$$

When the pointer is at the zero of the capacitance scale the capacitance of the tubular portion is 14·11 $\mu\mu$F, and since the length of the system is 18 cm the inductance of the tubular portion is 0·0255 μH. When the pointer is at 16 $\mu\mu$F on the capacitance scale the inductance is 0·012 μH, and so the obtainable change of inductance is 0·0135 μH. The inductance change may be found from the capacitance scale by means of the conversion chart in Fig. 16. Fig. 17 shows a resonance curve for a circuit whose inductance, when tuned to a frequency of 28·8 megacycles per sec., was 0·56 μH. Part of this was contributed by the tubular condenser used as an inductance. The resonance curve was determined by

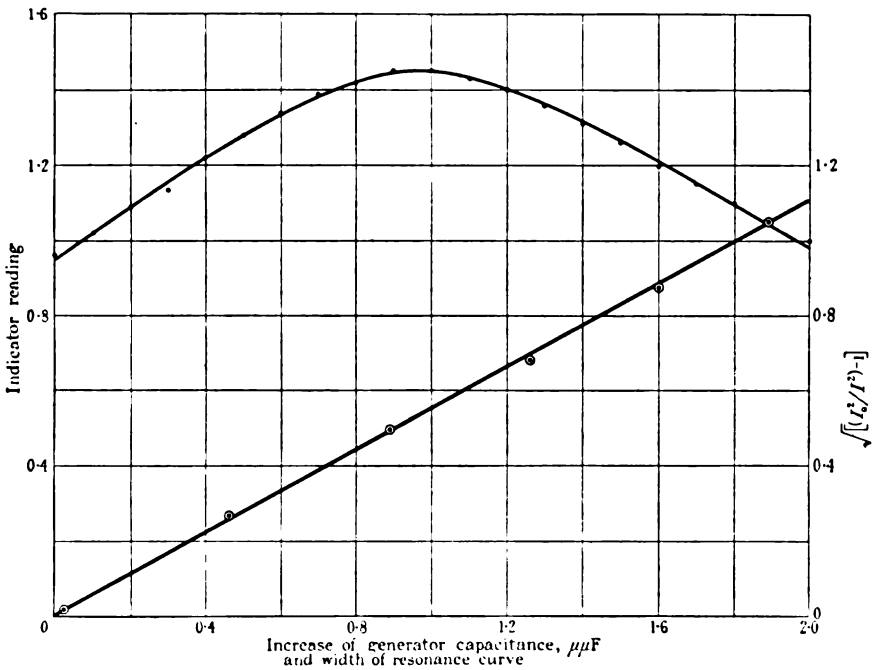

Fig. 15.

= 210 m. For receiver circuit, L = 70 μH. Generator capacitance = 50 $\mu\mu$F.

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plotting the readings of an indicator against the inductance change arrived at from the capacitance reading and the conversion chart of Fig. 16. The width of the resonance curve at 0·71 of its maximum height is found

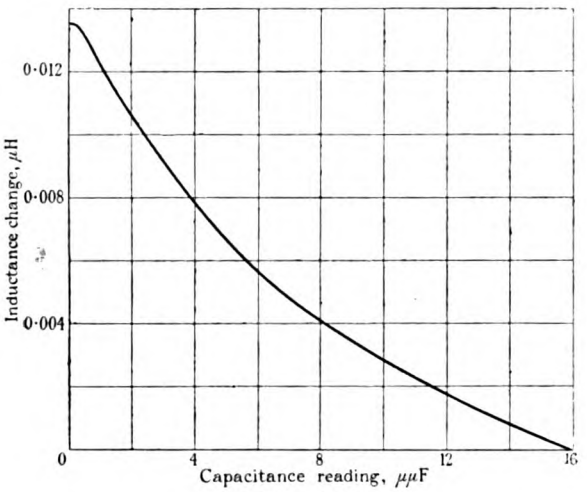

FIG. 16.

to be 0·0092 μH. It is interesting to see a resonance curve plotted within an inductance change of 0·01 μH.

(7) RESIDUAL INDUCTANCE OF THE CONDENSER.

One important use of the plunger switch is for determining the residual inductance of the condenser. By

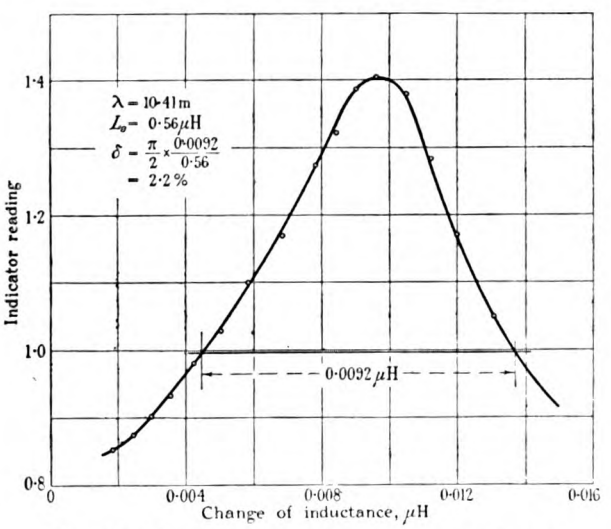

FIG. 17.—Resonance curve by inductance change.

connecting the condenser, short-circuited by the plunger switch, in series with an inductance of 0·98 μH, it was found by difference that the total internal inductance was 0·11 μH. When the instrument is used as a condenser the current will decrease almost uniformly from a maximum at the input terminal to zero at the top end of the rotor. Consequently the effective internal inductance will be about one-third of that measured with a uniform current between these two points and the residual inductance of the condenser is accordingly about 0·037 μH. The calculated inductance of the terminal tube is 0·045 μH and the calculated maximum inductance of the tubular portion is 0·025 μH, so that without end effect the value of the inductance when short-circuited would be 0·07 μH, which is of the same order as the measured value of 0·11 μH. Another way of estimating the residual inductance of the condenser is to add the calculated inductance of the terminal tube to one-third of the difference between the total measured inductance and the inductance of the terminal tube. This method gives the residual inductance as 0·056 μH, which must be an over-estimate. The residual inductance is probably somewhere between 0·04 and 0·05 μH, according to the value of the capacitance.

It would be troublesome to make similar measurements for a condenser with semicircular plates, because of the difficulty of arranging to take current from the circumference of the plates and so obtain a radial current-flow.

(8) APPLICATION TO A LOW-RANGE WAVEMETER FOR SHORT WAVES.

The tubular inductance and tubular condenser can be combined to form a robust and self-screening short-wave wavemeter of small range. One method of doing this is shown in the sectional drawing of Fig. 18. The circuit inductance is formed by a long rectangle of thick copper wire which is surrounded by a thick copper screen tube. A small loop at the end of the rectangle protrudes through the end of the screen tube and makes it possible for the wavemeter to absorb energy from the generator whose frequency is to be measured. The wavemeter capacitance is in two parts, one fixed and the other variable. The fixed portion is the capacitance between the screen tube and the outside surface of a copper cylinder concentric with the screen tube, and can be seen by reference to Fig. 18. The variable portion is formed by the capacitance between the inside surface of the concentric cylinder and the outside surface of a cylinder which revolves on an eccentric axle turning in bearings eccentric with the axis of the screen tube. Since the revolving cylinder is in electrical connection with the screen tube, it is at earth potential and consequently the dial handle need not be insulated. The whole of the capacitance is self-screening and all but a small fraction of the inductance is screened, so that the natural frequency of the circuit cannot be disturbed by surrounding objects. Since the whole wavemeter is enclosed in a thick copper tube the instrument is extremely robust and is suited to portable and field use. When suitably designed it must also be very permanent in calibration because small mechanical displacements make very small changes of capacitance or inductance.

If two thin wires, each of radius r, are placed symmetrically inside a screen tube of radius a and if each wire is at a distance b from the axis of the tube, it is readily shown that the wires will have an inductance per unit length given by the formula

$$L = 4 \log_e \frac{2b}{r} \cdot \frac{a^2 - b^2}{a^2 + b^2}$$

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For a given value of r this is a maximum when $b = 0{\cdot}49a$, and hence

$$L_{max.} = 4\log_e(1{\cdot}224b/r) = 4\log_e(6a/r)$$
$$= 0{\cdot}921\log_{10}(6a/r)\ \mu\text{H per metre}$$

The inductance changes slowly with the distance between the wires, because it is the flux close to the wire which provides the dominant fraction of the energy in the field. The inductance varies as the logarithm of a ratio and this changes slowly with the ratio. For example, suppose two wires, not in a screen tube, are spaced 20 radii apart; if the spacing is increased to 25 radii the inductance will be increased by only 7 per cent. The small effect of linear displacement may be illustrated by a numerical example. If the wires are of 1-mm radius and placed inside a tube of 30-mm radius they should be spaced 29·4 mm apart in order to obtain the maximum inductance. It may be shown, however, that if some accidental displacement moves them to a distance of 31·4 mm apart, the extra spread of 2 mm (6·7 per cent) increases the inductance by 0·4 per cent. Much greater immunity from change would be obtained byusing a smaller wire or a larger screen tube.

It may be shown that the resistance per unit length of the system is given by the formula

$$R = \left[\frac{8ab^2}{a^4 - b^4} + \frac{b}{r\sqrt{(b^2 - r^2)}}\right]\sqrt{(f\rho)}$$
$$\doteqdot \left(\frac{1 + 2{\cdot}05r}{a}\right)\frac{\sqrt{(f\rho)}}{r}, \text{ when } b = 0{\cdot}49a \text{ and } b >> r$$

Thus if (a/r) is of the order of 30 the eddy currents in the screen increase the effective resistance by about 7 per cent.

If the screen losses are ignored it may be found that the decrement of such an inductance of copper wire can be calculated from the formula

$$\delta = \frac{1}{150r\sqrt{(f)}\log_{10}(6a/r)}$$

where f is the frequency in megacycles per sec. and r is in cm. If $r = 1$ mm, $a/r = 30$ and $f = 30$ megacycles per sec. ($\lambda = 10$ m) it may be found that $\delta = 1$ per cent, and this will be increased to about 1·07 per cent by the losses in the screen tube. So far as copper losses are concerned, the wavemeter will have a very small decrement and therefore the resonance condition can be accurately determined.

The discussion in the earlier portions of this paper shows that a small accidental eccentric displacement of the condenser cylinder will cause a very small change of capacitance, provided that the radius of the latter is considerably smaller than that of the screen tube. The ratio of the radii and their actual values should be so chosen that some assigned movement, considered as the upper limit possible, cannot alter the capacitance by more than a fraction which will be negligible for the particular standard of accuracy laid down. The eccentricity of the inside rotating cylinder must be arranged to give the required range of wavelength—which may be, say,

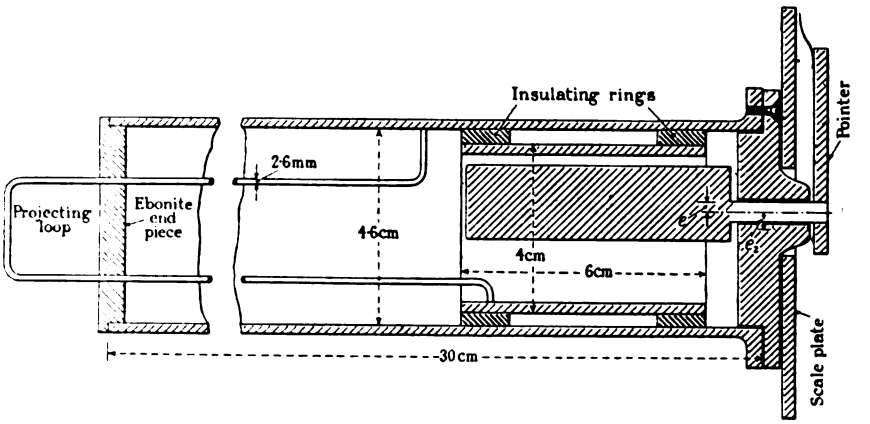

FIG. 18.

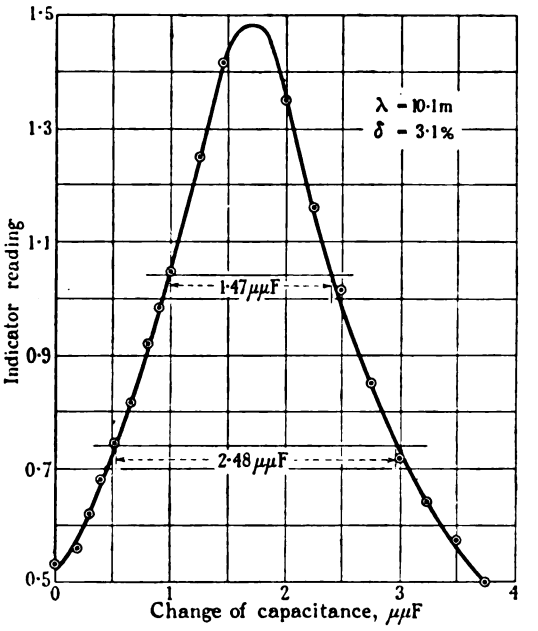

FIG. 19.

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± 1·5 per cent. The dial should be calibrated in terms of capacitance change as well as in terms of wavelength, and it will then be possible to plot a resonance curve for the finished instrument and so determine its decrement. With suitable design and careful manufacture it may be possible to make a wavemeter of this form which will have a temperature coefficient of the same order as the temperature coefficient of linear expansion of copper.

The dimensions shown in Fig. 18 are not the most suitable, and they were dictated by the material available. The instrument shown in Fig. 18 was used with a thermionic voltmeter as indicator, and this was connected across a few centimetres of the earthy leg of the inductance. In a standard instrument it would be preferable to connect the inductance to the screen tube through the heater of a thermal ammeter. The instrument shown in Fig. 18 has a wavelength range from 10·0 to 10·4 metres and the range of capacitance is about 8 $\mu\mu$F. Fig. 19 shows a resonance curve for the complete instrument and this was obtained by using the calibrations of the variable portion of its own condenser. The decrement of the complete instrument is about 3·1 per cent. A maladjustment of 0·15 $\mu\mu$F, corresponding to about 1·7 degrees of arc on the steepest part of the scale, can easily be detected. This means that the instrument is capable of measuring a wavelength correct to about 0·07 per cent.

Much greater accuracy of comparison can be obtained if the resonance point is determined by measuring the capacitance change required to move across the resonance curve at any assigned portion of its maximum height.

In order to increase the inductance without appreciable increase of bulk, there is no objection to enclosing four or more wires in the screen tube.

Discussion before the Wireless Section, 7th January, 1931, on the Papers by Messrs. Colebrook and Wilmotte (see page 49), and Mr. Moullin (see page 59), respectively.

Mr. P. W. Willans: My principal experience of the use of Messrs. Colebrook and Wilmotte's method of measuring high-frequency resistance was gained in the spring of 1926, while working in the Igranic Co.'s laboratories. We encountered exactly the same troubles as are described in the paper, and arrived at precisely the same method of working. As we found that it was necessary to employ a constant frequency when observing the two stationary points of frequency, we adopted the method which the authors describe of injecting an audio-frequency voltage after the detector and observing the beat frequency between it and the heterodyne note. This practice was also varied somewhat by employing a third radio-frequency oscillator set to the frequency at which it was desired to make observations, beats being obtained between it and the oscillator the frequency of which was being varied by the coupled circuit. In the course of measuring a whole range of commercial coils it happened that in the case of some long-wave coils we were unable to obtain a fine enough observation, even with the aid of beats, and we therefore made use of harmonics in the way described by the authors. The method was not employed at narrow limits of accuracy as we were engaged in revising the gauges of wire used in these commercial coils so as to obtain the best high-frequency resistance, and a rough, quick observation was all that was required. I am chiefly interested in the question of the oscillator to which the circuit under measurement is coupled, and I appreciate the authors' remarks regarding the variation of the frequency of this oscillator with resistance. Although I cannot speak from definite experience, I feel that the trouble is essentially a variation with amplitude, and that the frequency change is a result of a change in the amount of grid current flowing. From some work which I once carried out on the constancy of frequency of an oscillator, it appeared that the flow of grid current through the leakage inductance between oscillator and reaction coils gave a quadrature component of e.m.f., and a consequent distortion of frequency. If means are adopted of ensuring exact phase opposition between the e.m.f.'s on plate and grid, frequency distortion of this type disappears. One way of carrying this out is by tuning out the leakage inductance with a condenser, but a far better method consists in closely coupling the two coils. A grid-tuned oscillator was constructed on these lines in which the anode winding consisted of a number of sections wound coaxially with the main solenoid but spaced away from it, these sections being connected in parallel. In this way the mutual inductance was kept to a sufficiently low value for steady oscillation of good wave-form, and at the same time the coefficient of coupling was brought near to unity. It was found that an oscillator of this type had good frequency stability for variations of amplitude, and it was quite satisfactory when used for resistance measurements by the present method. Another variant of this form of oscillator consists in connecting the reaction coil, wound as described above, in the tuned circuit. This still further increases the effective coupling between the two windings and promotes opposition of phase between plate and grid e.m.f.'s. It has the inconvenience, however, that both plates of the tuning condenser are at live potential. Mr. Colebrook's proposed dynatron oscillator evades all of these difficulties and is probably the best arrangement, provided that screen-grid valves of sufficient stability are available. The authors refer to the discontinuous variation of frequency of an oscillator to which a tuned circuit is too closely coupled, and in this connection I am curious to know what is the limit of close coupling for satisfactory resistance measurement. It is clear that the two turning points of frequency can be observed even under conditions of coupling such that a click is produced in passing from one to the other. What I should like to know is whether the method is valid in these circumstances. On the face of it, the closer the coupling the greater the sensitivity, but it would appear possible that at very close couplings errors of a secondary nature might creep in and that in consequence it would be safer to couple more loosely and observe the frequency change more carefully. The authors' proposals for measuring impedances generally by means of this method are of great interest, and I hope we shall hear of developments in this connection.

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Mr. P. K. Turner: Mr. Moullin remarks on page 63 that formula (1) of his paper "is troublesome for numerical calculation, and it involves retaining a great many significant figures." I do not see why this should be so, for this formula is expressed in terms of an item β, which the expression on the next line reveals at once as the cosine of an angle (real or imaginary). By the ordinary methods for the solution of triangles the sine of half that angle may easily be found from a, b and d in a way which, unlike the direct calculation, does not involve the algebraic difference between two squares. When β has been evaluated in this way the denominator of formula (1) may be treated as an inverse hyperbolic function of β. Will it not thus be possible to obtain the value of c correct to 4 significant figures from 4-figure tables, or to 5 figures from 5-figure tables?

With regard to the paper by Messrs. Colebrook and Wilmotte, despite the fact that I published the method described I do not believe that it is as accurate as other methods, given equal accuracy of apparatus. The simple resistance-variation method, for example, is based upon a formula involving differences of resistance which are large in comparison with the resistances themselves. Thus if the resistances are known correct to one part in a thousand, the other items in the formula will be roughly of the same order of accuracy. This method necessitates a measurement of voltage also, and the resistance finally obtained is of the same order of accuracy as that of the values of resistance and voltage to be measured. In the method described in the paper, as also in the ordinary resistance-variation method, it is necessary to know the frequency to the same order of accuracy as that which is desired in the final result. Such accuracy is easily obtained in practice. The capacitance, however, has to be known to a much higher order of accuracy. According to the authors' statement on page 53, the capacitance changes can be repeated with a maximum variation of 0·05 $\mu\mu$F, which corresponds to an accuracy of the order of 0·25 to 3·0 per cent. Roughly speaking, therefore, it is necessary to know the capacitance to at least one hundred times greater accuracy than can be expected in the final result. Even with Mr. Moullin's new condenser this would be an extremely difficult achievement. The total capacitance change measured, according to some of the examples in Table 1, is of the order of 3 to 6 $\mu\mu$F, except where additional resistance has been inserted. Differences of this magnitude have to be observed in a total capacitance of 500 $\mu\mu$F, and the change must be measured to an accuracy within 1 or 2 per cent. When this method first came to my notice I stressed its advantages, for at that time I was writing chiefly for radio amateurs who had no current- or voltage-measuring devices suitable for radio frequencies. Where only a calibrated condenser is available and an accuracy of 3 or 4 per cent in the results is sufficient the Willans method surpasses all others. Nowadays, however, one is more likely to possess an accurate valve voltmeter than a condenser of the accuracy required for this measurement. I have frequently been able to determine radio-frequency resistances to an apparent accuracy of 0·25 to 0·5 per cent by resistance-variation measurements, and I should be sorry to have to pay for the condensers which would be required in order to work to the same accuracy by the authors' method.

Mr. B. Williams: I have frequently used the method described by Messrs. Colebrook and Wilmotte for the accurate determination of the resonance point in absorption wavemeters. It is excellent for quick calibration of a number of absorption wavemeters, particularly at very high frequencies. The method I have adopted is slightly different from the authors' but very similar to the one referred to by Mr. Willans. I do not use the audio-frequency source. Two radio-frequency oscillators of good frequency stability are employed, one being coupled to the test circuit. The heterodyne notes produced by each of these oscillators in a third oscillator (usually a simple self-oscillating receiver, the frequency stability of which is unimportant) are made to "double beat" by adjustment of one of them. The advantage is that the listening apparatus does not form part of the measurement circuit, and the frequency of the heterodyne note can be anywhere in the audible range. The method is used chiefly to determine the dynamic resistance $(L^2\omega^2)/R$ of coils, which, by a simple manipulation of the formula, can be shown to be equal to $1/(\omega dc)$, where dc = change in capacitance from the resonance point to either peak. This capacitance change can be very accurately determined. I am sorry that no results of impedance measurements by this method, with the exception of the one for a grid-leak, are given in the paper. When attempting to use the method for determining the input impedance of screen-grid valves under operating conditions in which it was not permissible for the applied radio-frequency voltage to exceed about 0·2 volt, it was found that in order to obtain any results at all the voltage injected into the test circuit across which the screen-grid valve was connected had to be of such an order that grid currents were immediately produced and all the results were vitiated. Could the authors give any information on this point? Some of the figures in Table 1 appear to me to be incorrect, because the changes in capacitance do not agree with the apparent values of the resistance.

With regard to Mr. Moullin's paper, it is possible that the checking of the calculated calibration against a standard condenser could have been done with more precision by means of the double-beat method, which, in my experience, is much more accurate than the ordinary resonance method using a valve voltmeter. With regard to the new type of condenser, the only disadvantage of these instruments is that they take so long to make. I have now had one on order for about 6 months.

Mr. F. J. H. Charman: I worked with Mr. Willans when he was making the investigations to which he has referred; one of the first difficulties we encountered was that of obtaining enough frequency change without getting too tight a coupling. The coils which were then available were not quite so good as those exhibited by the authors. They had power factors of 0·01, 0·03, and 0·4, and below about one million cycles it was difficult to obtain an appreciable change of frequency as we tuned the test circuit. The difficulty was overcome by using the harmonics of frequencies below one million cycles. The frequency change was then mul-

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tiplied, and, instead of trying to work with a frequency change of, say, 1 000 cycles, we could, if we wished, get a change of 10 000 cycles by using the 10th harmonic. If the coupling was made too tight the resistance of the tuned circuit was apparently raised, the distance between the peaks widening as the coupling increased until a discontinuity occurred in the frequency curve of the oscillator. (The oscillator was weakly excited, to secure good wave-form.) One could therefore always make sure that the coupling was not too tight by observing any variation in width of the peaks produced in the oscillator frequency when the coupling was varied. The accuracy required is not quite so high as has been suggested by Mr. P. K. Turner. What we really want to know is the change in capacitance, and not the total capacitances, so that a condenser which will cover 2 or 3 $\mu\mu$F will enable us to measure a change in the total capacitance to quite a small fraction of a micromicrofarad without much difficulty. One of the chief advantages of this method of measuring resistances is that it does not involve interfering with the test circuit. There are cases where, as on short wavelengths, it is impossible to use other methods, such as that of inserting resistances into the circuit. Nobody has yet produced high-frequency resistances of known value which, when inserted into the circuit, can be relied upon at frequencies of the order of 10 megacycles. I have tried to use this method on short waves, and under such conditions I believe that it is necessary to be careful about capacitance coupling. I know of no other method applicable to short waves which is comparable with the one described by the authors.

Mr. A. Wilkinson: The chief advantages of the method described by Messrs. Colebrook and Wilmotte are its simplicity and the fact that it gives fairly good results with little expense beyond that involved in the provision of good-quality condensers. The remainder of the apparatus necessary is such as can be found in any ordinarily well-equipped wireless laboratory.

The condenser described in the paper by Mr. Moullin promises to be a useful instrument, for although many types of good condensers of the order of 100 $\mu\mu$F and upwards are available, there are few on the market which are accurate on the low-range scales. For example, I have two condensers which are supposed to have a maximum capacitance of 10 $\mu\mu$F, but they function far more like inductances than condensers. I should like to ask Mr. Moullin whether the choice of mycalex for the insulating bushes is final. Mycalex has a relatively high power factor, and I think that the condenser would be improved if silica were used for the bushes instead. I suggest that in a condenser of such high quality the additional condensers should have air rather than mica as their dielectric. Two small air condensers would be more useful, more permanent, and give greater confidence in the performance of the instrument than the mica condensers mentioned in the paper.

Mr. L. B. Turner: The method described by Messrs. Colebrook and Wilmotte appeals to me especially because of its suitability for very short waves, and for this reason it may benefit from the condensers described in the paper by Mr. Moullin. While not questioning, so far as long or medium-length waves are concerned, what Mr. P. K. Turner has said about the excellence of the inserted-resistance method, I do not think this can be conveniently applied at wavelengths such as 20 m. For example, I do not see that it would be practicable to measure the resistance of a half-wave antenna by cutting it in two and inserting a known resistance. The authors refer to the inconstancy of oscillators as a possible practical difficulty in the use of the method, but this strikes me as unnecessarily fearful. The only relevant inconstancy seems to be variation of the frequency of the auxiliary acoustic oscillator; but the method only requires that this frequency shall be fixed during an observation, taking perhaps 1 or 2 minutes, and the oscillator is not interfered with during the experiment. Inconstancy of the high-frequency oscillator does not matter, as the frequency of the latter is adjusted during the observation by the acoustic note.

With reference to Mr. Moullin's paper, the author, when proposing to use the condenser as a variable inductance, treats the instrument as a transmission line, and argues that because he knows the capacitance he knows the inductance as well. Does not this argument require that the condenser should behave like a transmission line with uniformly distributed capacitance and inductance? It would seem, however, that the paths of the current along the rotor, at least in its complicated tracks along the ribs and vanes from the axle to the cylindrical surface, makes the transmission line a very unreal equivalent. What current, for example, is it which has the length $l = 18$ cm referred to on page 67? I shall be glad if he will explain why, in determining the inductance of the condenser, it is unnecessary to take into account the path by which the current passes through the instrument.

Dr. L. E. C. Hughes: I think that many of the mathematical expressions given in Mr. Moullin's paper are not so well known as is implied by the author, and it would be useful if references to these were given or if some of the expressions were worked out in appendices. With regard to Mr. Moullin's condenser, Fig. 6 shows a lead-in to the inner cylinder by a small spiral of wire. Such a spiral has a small inductance, which is important in some other forms of condenser. In the present instance the spiral is short-circuited by the rubbing contact, and it would therefore appear to be ineffective. The question of the distribution of the charging current does not at present arise, because the dimensions are small in comparison with the wavelength.

Mr. A. J. Gill (*communicated*): The condenser described by Mr. Moullin is an ingenious development, and one which is to be recommended on account of its possibilities of mechanical rigidity, electrical constancy, perfect screening and ease of conversion to an inductance. Practically the only undesirable feature is the irregular scale. Such an irregularity might be advantageous if the minimum value of capacitance were very low in comparison with the maximum; this condition, however, does not obtain. As a result there is a widely varying accuracy throughout the range, and this makes it less suitable for use in a wavemeter. About two years ago I had to produce a small panel-mounting condenser having a total capacitance of about 16 $\mu\mu$F which

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should be capable of being read to about 0·002 $\mu\mu$F, for use in precision measurements of short wavelengths. As this condenser was to be used in parallel with a larger one it was essential that it should have an approximately linear scale. The design adopted is shown in Fig. A. Movement is produced by means of a 1-in. micrometer head reading to 1/10 000 in. The vertical assembly avoids the use of return springs. In its lowest position the moving electrode remains within the upper cylinder, which is sufficiently long to render end-effects negligible.

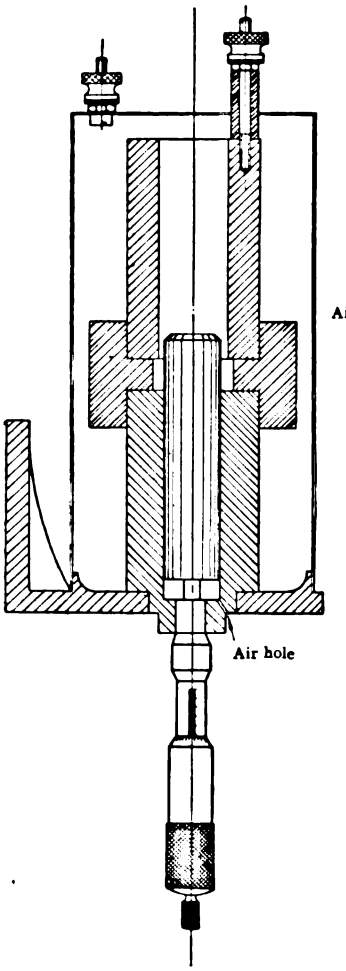

FIG. A.

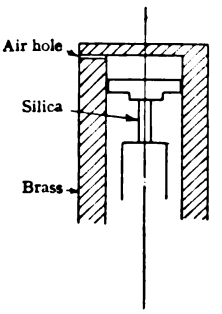

FIG. B.

As a result the scale is linear to within a fraction of 1 per cent. After two years' daily use this condenser still continues to be entirely satisfactory. Fig. B shows a possible modification to the inner electrode, in order to permit shortening of the upper cylinder without sacrificing linearity.

Mr. F. M. Colebrook (*in reply*, on behalf of himself and Mr. R. M. Wilmotte): I am interested in Mr. Willans's remarks about frequency stability in oscillators and the means he adopted for securing this stability. I have prepared a brief paper on the general conditions of frequency stability in valve-maintained oscillators, indicating certain possible lines of development which it is hoped to investigate in the future. In the meantime the dynatron oscillators used in the present form of the apparatus* are stable enough to be free from error due to frequency drift in the course of the measurements. This, as Mr. L. B. Turner points out, is not a very exacting requirement. With regard to the limit of permissible steepness of the centre part of the S-shaped curve, this is a matter for actual trial and I have not investigated how far one may safely go in this direction. It should be pointed out, however, that the error associated with the frequency-resistance characteristic of the oscillator is only independent of coupling as long as the characteristic is linear. Too close a coupling might cause a departure from this condition, so that the increased sensitivity might be paid for in decreased accuracy. Incidentally, I am told by Prof. Appleton that the discontinuity of frequency variation associated with too close a coupling is due to the fact that the centre of the S-shaped curve actually tilts forward after becoming vertical, giving three possible frequencies for a given tuning capacitance of the first or measurement circuit. The discontinuity is thus a jump from one possible frequency to another. The application of the method to the measurement of impedance will be illustrated in a paper now in course of preparation, dealing with the behaviour of receiving aerials.

Mr. P. K. Turner evidently prefers the resistance-variation method. I suggest that this may be partly due to the fact that he is much more familiar with it than with the method now described. The resistance-variation method gives a result in terms of other resistances, the values of which at the frequency in question have either to be measured by some other method or to be determined as far as possible by calculation based on their d.c. value. The present method, on the other hand, is an absolute one, giving resistance expressed in terms of frequency and capacitance. It should also be noted that the inserted-resistance method is subject to a small but appreciable error on account of the self-capacitance of the coil (see footnote on page 54). However, each method has its own merits and demerits and there is plenty of room for both.

Mr. Williams's idea of using an additional radio-frequency oscillator instead of the audio-frequency oscillator is an interesting and useful one, at least as far as resonance detection is concerned. As applied to the actual resistance measurement it might lead to some confusion of beat notes, but if Mr. Williams has actually used it in this way it would seem to be manageable. I have not actually tried it but will certainly do so. The apparent arithmetical errors to which he refers are due to miscopyings in the C_1 column. The errors have been corrected in the table as printed in the *Journal*. As far as the measurement of impedance is concerned, as stated above the method has been successfully used for an investigation of aerial impedance. The difficulty Mr. Williams mentions in connection with the measurement of the input impedances of valves is a serious one and has yet to be overcome. It is possible that the method may be found impracticable for this purpose, but the possibilities have not yet been fully explored.

The comments made by Mr. L. B. Turner, Mr. Wilkin-

* See Note, page 58.

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son and Mr. Charman are favourable and do not call for any reply beyond an expression of appreciation.

I should like to make it clear that it has not been practicable to refer the points raised in the discussion to Mr. Wilmotte, who is in America.

Mr. E. B. Moullin (*in reply*): Mr. P. K. Turner suggests that formula (1) of my paper should not be troublesome to evaluate numerically. I did not wish to imply that I had used the clumsy process of direct numerical substitution. It is first necessary to express d as a function of θ, after which, formula (1) can be changed to a more elegant form. Even when this has been done, however, the computation is very laborious. As the calibration was to be performed entirely by calculation, I thought it desirable to determine the value of θ, corresponding to a given capacitance, to an accuracy appreciably greater than that to which the pointer of a dividing engine could be set. Any errors in dividing the scale would then be due only to the dividing machine, and not partly to the calculation.

Dr. Hughes will find formula (1) derived in Dr. A. Russell's "Alternating Currents" (vol. 1), and formula (2) derived in article 329 of Sir James Jeans's "Electricity and Magnetism": they can also be found in many other well-known textbooks.

I agree with Mr. Williams's suggestion that the double-beat method would have been preferable for checking the calibration against the standard. It seemed unnecessary in this case, because undoubtedly the main source of error was in setting or in reading the pointer of the standard condenser. Fig. 10 shows that the law of the scale, and its values, appear to agree with a small standard which was believed to have a uniform scale. Additional accuracy of comparison would, I think, merely have served to show imperfections in setting the standard.

Mr. Wilkinson queries the use of mycalex for the bushings. Mycalex is quite suitable if the condenser is to be employed only at radio frequencies, for the volume of mycalex used is very small and the power factor of this material is about 0·17 at 100 kilocycles and 0·13 at 1 000 kilocycles. For the whole condenser, therefore, the resultant power factor introduced by the mycalex must be very small. Fused silica would, no doubt, be better, but is it worth the extra trouble and expense? The choice between air and mica condensers for the base chamber is not so obvious as it may seem. Resonance curves have been plotted for the same circuit when the air portion only is used and also when a base condenser plus a small fraction of the air portion is employed. No difference can be detected in the circuit decrement whether the capacitance is all air or mainly mica: this shows that the air condenser and the mica condenser have the same power factor within a very small fraction of 1 per cent, so that it does not seem that there is much advantage to be gained by having air condensers in the base. Air condensers having a capacitance of 15 $\mu\mu$F are not very easy to construct rigidly, and have at the same time a high insulation resistance: what is there to be gained by using them? The real difficulty is due, not to dielectric loss but to direct leakage. At 50 cycles per sec. the impedance of the condenser is 200 megohms, so that an insulation resistance of 2 000 megohms would give a power factor of 10 per cent to the condenser: at 5 megacycles per sec. the same leakage would give a power factor of 10^{-4}. If the condenser is to be used in a low-frequency bridge, it is insulation resistance which matters and not dielectric loss. In my opinion it is easier to make a small mica condenser perfectly insulated than a small air condenser. If the condenser is to be used in a low-frequency bridge, then I think it ought to have silica bushings; but if it is for use at radio frequencies only, silica appears to be unnecessary.

Mr. L. B. Turner has queried the applicability of the transmission-line formula, and in this he has some justification. Along the major portion of the rotor and the input lead, the inductance is uniformly distributed, but at the junction of the terminal lead and the rotor there is a discontinuity, the general effect of which will be to increase the effective length of the system. Even so, the effective inductance cannot be appreciable until the wavelength is well below, say, 10 m. The advantage of the tubular system is that the residual inductance can be estimated. If the wavelength is so short that the correction is important, a shorter instrument should be used. It might be better to fit a top and bottom cover to the rotor so that the current would not have to wander along the ribs and vanes. This would slightly increase the minimum capacitance, and decrease slightly the residual inductance. The system does, however, permit an inductance measurement to be made, and the value so found must be considerably greater than the effective residual inductance of the condenser. So, whatever the residual inductance may be, it is certainly less than 0·1 μH, and therefore it is always possible to place an upper limit to the inductive effect. Since experiments show that the change in capacitance agrees with the calculated values, it seems certain that the change in inductance will also do so, and the resonance curve of Fig. 17 appears to substantiate this.

I am very interested in the design of Mr. Gill's condenser. He has commented on the non-uniformity of the scale of my instrument, which would certainly be more desirable for some purposes if the scale were more uniform. It could, I think, be very much improved by suitably shaping the rotor. Now that I have proved that the calibration of the circular cylinder can be calculated, I propose to make instruments with rotors so shaped as to give a more uniform scale. Such instruments can then be calibrated experimentally from one made to the original design.

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SOME MEASUREMENTS ON A LOUD-SPEAKER IN VACUO.*

By P. K. TURNER, Member.

(*Paper first received 29th November, 1930, and in final form 19th January, 1931; read before the* WIRELESS SECTION *4th February, 1931.*)

SUMMARY.

The electrical impedance of a moving-coil loud-speaker is measured under three different conditions: (1) with the coil held fast; (2) with the coil free, but *in vacuo*; (3) under normal conditions.

From the basic analysis of this type of instrument, which is given briefly, it follows that this series of measurements should allow a sufficient separation of its various characteristics to lead to the possibility of estimating its actual performance, and also to indicate whether any defects shown are electrical, mechanical or acoustic in origin.

The measuring equipment is described briefly, and the results are given of a series of measurements between 40 and 6000 cycles per sec. From these the characteristics of the instrument are derived.

The results give some indication as to why loud-speakers of this type can give a much better performance at high frequencies than would be expected from simple theory: the inductance of the coil and the mass of the diaphragm, which both cut down the output at high frequencies, are shown to fall off rapidly as the frequency rises, while the acoustic resistance does not fall off to the same extent and is, in fact, much in excess of the theoretical radiation resistance.

The characteristics found are used to re-compute the input impedance, and give good agreement with the original observations. The computation is extrapolated to obtain estimates of the output and efficiency at frequencies for which the actual observations do not give opportunity.

LIST OF SYMBOLS.†

(1) *In Section 5(c).*

α = propagation constant of line.
C = distributed capacitance of a line (electrical).
C_1, C_2 = capacitances in the electrical network equivalent to the moving parts of a loud-speaker.
c = mechanical compliance of the cone.
G = leakance of the electrical equivalent line.
g = mechanical "leakance" or internal dissipation of cone.
L = distributed inductance of the equivalent line.
L_1 = a concentrated inductance in the electrical equivalent network.
m = mass of cone.
m' = apparent mass of cone.
R = distributed series resistance of equivalent line.
R_1, R_2, R_3 = concentrated resistances in the electric equivalent network.
R' = mechanical input resistance of the cone as a line.
Z_0 = surge impedance of (mechanical or electrical) line.
Z' = input impedance of line.

(2) *Apart from Section 5(c).*

A = "force factor," or electro-mechanical mutual impedance, C.G.S. units. (2 *b*).
B = effective flux-density in gap, gauss. (2 *b*).
c = mechanical compliance of cone itself, cm/dyne. (5 *c*).
C_p = motional capacitance ($= Y_p/\omega$), farads. (6 *b*).
γ = coupling factor in eddy-current effect. (6 *a*).
D = electro-mechanical conversion factor ($= A^2 \times 10^{-9}$). (2 *b*).
δ = reciprocal of time-constant of eddy-current path. (6 *a*).
E = e.m.f., volts. (2 *b*).
f = frequency, cycles per sec. (2 *a*).
F = mechanical force, dynes. (2 *b*).
g = mechanical leakance of cone. (5 *c*).
G_p = conductive component of $1/Z_r$, mhos. (2 *a*).
η = output ratio. (8).
I = current, amperes. (2 *b*).
k = R/X or X/R in series-parallel conversions. (6 *b*).
l = length of wire in coil, cm. (2 *b*).
L = inductance of coil *in situ*, henrys. (5 *b*).
L_4 = inductance of coil *in situ*, in absence of eddy-current effects, henrys. (6 *a*).
L_5 = inductance of eddy-current path, henrys. (6 *a*).
m = mass of moving parts, grammes. (2 *a*).
m_a = "acoustic mass," grammes. (7 *a*).
m' = apparent input mass of moving parts, grammes. (6 *c*).
M = mutual inductance of coil and eddy-current path, henrys. (6 *a*).
μ = coupling coefficient of above ($= M/\sqrt{[L_4 L_5]}$). (6 *a*).
P_{cv} = acoustic pressure for constant-voltage input (arbitrary units). (10).
P_{ci} = acoustic pressure for constant-current input (arbitrary units). (10).
q = $k + 1/k$. (6 *b*).
r = projected radius of cone. (2 *a*).
R = loud-speaker electrical resistance in general. (2 *b*).
R_1 = loud-speaker electrical resistance: "held." (3 *b*).
R_2 = loud-speaker electrical resistance: "normal." (2 *b*).
R_3 = loud-speaker electrical resistance: "vacuum." (2 *c*).
R_4 = resistance of coil in absence of eddy-current effects. (6 *a*).
R_5 = resistance of eddy-current path. (6 *a*).
(ohms)
R_a = acoustic resistance, C.G.S. units. (2 *a*).
R_a' = theoretical acoustic resistance of rigid disc, C.G.S. units. (7 *b*).
R_c = electrical resistance of coil, ohms. (2 *b*).

* Reprinted from *Journal I.E.E.*, 1931, vol. 69, p. 591.

† The references in brackets are to the sections of the paper in which the respective symbols first occur.

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R_m = mechanical resistance of diaphragm, C.G.S. units. (2 *a*).
R_d = total mechanical and acoustic resistance of diaphragm, C.G.S. units. (2 *a*).
R_{d_2} = total resistance of diaphragm, for "normal" case. (7 *a*).
R_{d_3} = total resistance of diaphragm, for "vacuum" case. (7 *a*).
R_p = electric "motional" resistance, as part of a parallel circuit. (2 *b*).
R_r = electric "motional" resistance, as part of a series circuit. (2 *b*).
R_u = useful acoustic resistance. (7 *b*).
s = suspension stiffness, dynes/cm. (2 *a*).
S = projected area of cone. (2 *a*).
U = displacement, cm. (10).
v = velocity, cm/sec. (2 *b*).
W_1 = electrical power input to speaker, watts. (10).
W_2 = acoustic power output from speaker, watts. (10).
X = (With the following exceptions, X with any suffix corresponds to R with the same suffix, as given above.)
X'_d = mechanical input reactance of cone, in general, C.G.S. units. (7 *a*).
X'_{d_2} = mechanical input reactance of cone, for "normal" case, C.G.S. units. (7 *a*).
X'_{d_3} = mechanical input reactance of cone, for "vacuum" case, C.G.S. units. (7 *a*).
Y_p = "motional" reactance ($= \omega C_p$). (2 *b*).
Z = (In all cases, $Z_h = R_h + jX_h$, for any value of h).
ω = pulsatance ($= 2\pi f$), radians per sec. (2 *a*).

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(1) INTRODUCTION.

(*a*) *General.*

The extraordinary difficulty of interpreting the results obtained by any measurements of the actual acoustic output of a loud-speaker is common knowledge among those who have done much work on the subject, though perhaps not yet fully realized by the general body of acoustic engineers. An excellent summing up is to be found in L. G. Bostwick's paper,* from which the following is a quotation:—

> "... From the above considerations it is obviously quite impossible to interpret the significance of response measurements on loud speakers in general unless such measurements are qualified by statements regarding the acoustic measuring conditions. Especially must information be given as to the position of the condenser transmitter relative to the loud speaker when the measurements were made, the method of measurement (pressure measured at one position or averaged within a region), and the size and nature of the medium. ... To determine which of two loud speaker response-frequency characteristics is the better involves, in addition to the above discussed acoustic considerations, an interpretation of the physiological significance of the magnitude and position in the frequency spectrum of departures in the curves from a straight horizontal line."

The whole of this paper is especially to be recommended in this connection.

On the other hand, pure analysis, even in the one case—the moving-coil speaker—which lends itself best to it, is not satisfactory except at quite low frequencies: it persistently indicates that the ordinary moving-coil speaker cannot be useful at frequencies over, say, $\omega = 5\,000$. For example, C. R. Cosens† arrives at outputs of 0·07 watt for 30 watt input at low frequencies, and 0·06 micro-watt at 4 000 cycles per sec., efficiencies of 0·25 per cent at the best, and response curves 20 decibels down at 4 000 cycles per sec.

Consideration of these and other difficulties led the author, in 1927, to the belief that the best line of attack was to check some of the fundamental data of the analysis by measurement: the measurement to consist in finding the input resistance and reactance of the instrument (at various frequencies) under three conditions. First, with the coil firmly held, so that the impedance of the coil itself *in situ* could be found. Second, with the whole instrument in a vacuum chamber: this gives an impedance depending only on (*a*) the coil impedance already found, and (*b*) the mechanical impedance of the moving parts, which latter can then be computed. Third, under normal conditions: here an acoustic impedance also is added, and can in turn be computed by allowing for the impedances already found.

By analysis of the results one should be able not only to compute the "response" curve of the speaker; but also to show whether any deviations from the ideal arise from electrical, mechanical, or acoustic defects. This latter information can in no way be obtained from a directly-measured response curve.

* See Bibliography, (1). † *Ibid.*, (2).

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Thanks to the enterprise of Mr. W. H. Lynas, then Managing Director of Graham Amplion, Ltd., and Mr. C. F. Phillips, chief engineer of the firm, who sanctioned the expenditure, the author was able to have a suitable vacuum chamber made, and towards the end of 1929 a relaxation in the pressure of other work enabled it to be put into use.

The first results are now presented, not so much for any intrinsic merit of their own (for they need confirmation by other workers), but because, in the author's opinion, they do indicate that this line of attack is a useful and powerful one.

(b) Some Summarized Results.

The measurements bring to light some interesting points, especially as to the means by which a reasonable output is obtained at high frequencies. Some of the more striking results may be summarized here:—

(1) The inductance of the coil falls very considerably as the frequency goes up, eventually becoming less than its value for the coil in air. This is due, apparently, to the " short-circuited secondary " effect of eddy currents in the iron in its neighbourhood. At low frequencies the effective permeability is of the order of 2.

(2) The effective mass (both mechanical and acoustic) of the moving parts also falls off very considerably as the frequency rises. In this case the apparent mechanical mass decreases from 21 grammes to the neighbourhood of 10 grammes for a frequency as low as 800 cycles per sec.

(3) The measured acoustic resistance does *not* fall off, and at all frequencies exceeds the theoretical radiation resistance of a rigid disc of the same diameter.

(4) Wave transmission down the cone becomes important at quite low frequencies, its effects being perceptible in this case at 300 cycles per sec.

(5) At low frequencies, about three-quarters of the width of the " surround " appears to behave as if it were part of the cone.

(2) The Basic Analysis.

(a) Diaphragm Impedance.

Let us first state shortly the fundamental equations of the moving-coil speaker.

If the moving parts have a mass m grammes, a stiffness s dynes per cm, and a resistance R_m dynes per cm per sec., we can put

$$Z_m = R_m + jX_m = R_m + j\left(\omega m - \frac{s}{\omega}\right) \quad . \quad . \quad (1)$$

If, in addition, there are an acoustic resistance and reactance R_a and X_a, we can put

$$Z_a = R_a + jX_a \quad . \quad . \quad . \quad . \quad . \quad (2)$$

and for the total mechanical impedance of the diaphragm

$$Z_d = Z_m + Z_a = R_d + jX_d = R_m + R_a + j\left(X_a + \omega m - \frac{s}{\omega}\right) \quad (3)$$

There are well-known equations, due to Rayleigh, for X_a and R_a as functions of diaphragm size and frequency. These are conveniently found in the excellent book by the late I. B. Crandall,* who also gives a simpler approximation. In Fig. 1 the author has given another approximation, the curves being for R_a/S and X_a/S against fr, where S = diaphragm area (cm^2), r = diaphragm radius (cm), and f = frequency (cycles per sec.).

These curves are for a flat disc in an infinite baffle: equations for a cone in a finite baffle are not yet available, so far as the author is aware. Further, these curves, also equations (1) to (3) above, are based on the assumption that the diaphragm moves as a rigid body. It obviously ceases to do this when, as the frequency rises, the wavelength of motion in the cone is no longer large compared with the size of the cone. We must then no longer assume $R_d = R_m + R_a$ and $X_d = X_a + \omega m - (s/\omega)$ in equation (3). R_d and X_d will in fact become the components of the input impedance of a transmission line. This matter will be taken up later.

(b) The Electro-Mechanical Conversion.

Any current I (amps.) in the coil will produce a force F (dynes) upon it, given by

$$F = A\frac{I}{10} \quad . \quad . \quad . \quad . \quad . \quad . \quad (4)$$

where A is called the " force factor," and is calculable by

$$A = Bl \quad . \quad . \quad . \quad . \quad . \quad (5)$$

where B is the mean flux-density (gauss) in the gap, and l is the length of wire (cm) in the coil.

Also, if the coil is in motion with velocity v (cm per sec.), it produces an e.m.f. E (volts), given by

$$E = A\frac{v}{10^8} \quad . \quad . \quad . \quad . \quad . \quad . \quad (6)$$

From equations (5) and (6) it follows at once that any mechanical impedance Z_d produces in the electrical circuit a reflected electrical impedance, Z_r, given by

$$Z_r = R_r + jX_r = \frac{D}{Z_d} \quad . \quad . \quad . \quad . \quad . \quad (7)$$

where

$$D = \frac{A^2}{10^9} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (8)$$

If now the coil has an electrical impedance of its own, $Z_c (= R_c + jX_c)$, then we have for the working input impedance of the speaker

$$Z = R + jX = R_c + R_r + j(X_c + X_r) \quad . \quad (9)$$

In analysing the results of our measurements, we start with (*a*) R_c and X_c, measured with the coil held fast, and (*b*) R_3 and X_3, measured with the coil free, but *in vacuo*. Simple subtraction gives R_r and X_r.

For the next step it is convenient to define some admittances.

Put

$$\frac{1}{Z_r} = G_p + jY_p \quad . \quad . \quad . \quad . \quad . \quad (10)$$

Then from equation (7)

$$\frac{Z_d}{D} = \frac{1}{Z_r} = G_p + jY_p \quad . \quad . \quad . \quad . \quad (11)$$

* See Bibliography, (3).

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To find G_p and Y_p, we have only to put

$$G_p = 1/R_p \quad . \quad . \quad . \quad . \quad . \quad . \quad (12)$$

$$Y_p = 1/X_p \quad . \quad . \quad . \quad . \quad . \quad . \quad (13)$$

to realize that R_p and X_p are the components which, in parallel, give the impedance Z_r.

From equations (3) and (11) we see that

$$R_d = DG_p, \; X_d = DY_p \quad . \quad . \quad . \quad . \quad (14)$$

and as in this case (*in vacuo*) there are no acoustic components, $R_m = R_d$, and $X_d = \omega m - (s/\omega)$, assuming rigid motion.

The length l of wire in the coil is known reasonably accurately, and its static value can hardly change for dynamic measurements. B, the effective flux-density in the gap, is also measurable statically, but it is by no means easy to arrive at it, for there is a very large fringing flux, and it is not at all simple to arrive at an estimate of the actual useful flux. s, the mechanical stiffness of the diaphragm suspension, can be measured fairly easily but not very accurately. The stiffness of the external suspension of the cone (the " surround ") must be excluded from the measurement. r, the cone radius, is easy to measure. m, the mass of moving

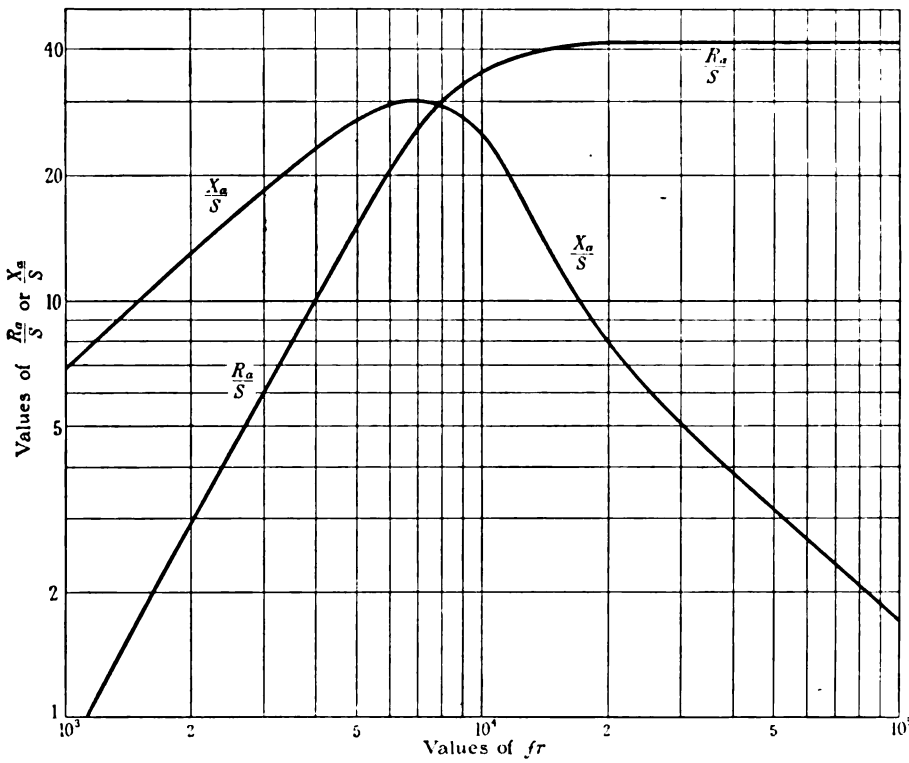

FIG. 1.—Acoustic resistance and reactance of a disc in an infinite baffle, as functions of frequency and diaphragm radius.

S = area of disc (cm²); r = radius (cm).

As will be seen later, the fact that m is known enables us to use our experimentally-derived value of X_d to find D and hence A; the value of s also follows.

From this point we repeat the analysis, but compare R_c and X_c with R_2 and X_2, the " normal " values for the speaker in air. New values for R_d and X_d are thus computed which differ from the previous ones by including also the acoustic components and hence allow us to compute the latter.

The exact form and scope of the computation will be detailed later in the paper.

(3) THE DATA.

(*a*) *Design Data.*

The data available before commencing the measurements proper are static values of r (hence S), m, s, l and B.

parts, is not so easy to measure as appears at first sight. Firstly, one has to consider as " moving " some part only of the elastic centering or suspension washer and the surround; secondly, the mass of the coil and its former, which is rigid except for very high frequencies (if carefully designed) must be separated from that of the cone itself, and this can only be done accurately by breaking up a finished sample, owing to the undetermined mass of adhesives, etc.

(*b*) *Observed Values.*

We have three sets of observed resistances and reactances, obtained under the following conditions:—

R_1 and X_1 (" held "). The coil and cone are not allowed to move. Actually, the speaker was placed horizontal, the coil supported by chocks in its central position, and about 2 kilograms of sand was then poured into the cone.

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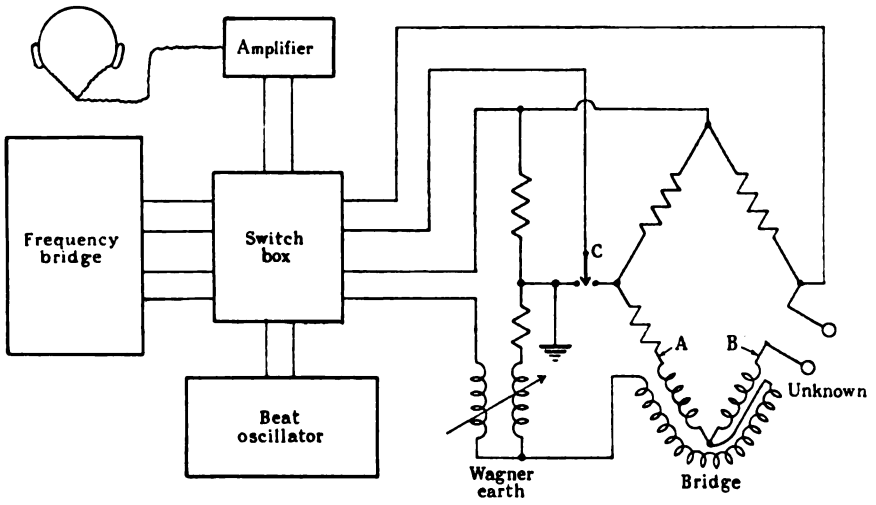

FIG. 2.—Simplified schematic diagram of the measuring equipment.

A and B are connection points for inductometer reversing switch. C is switch for Wagner earth.

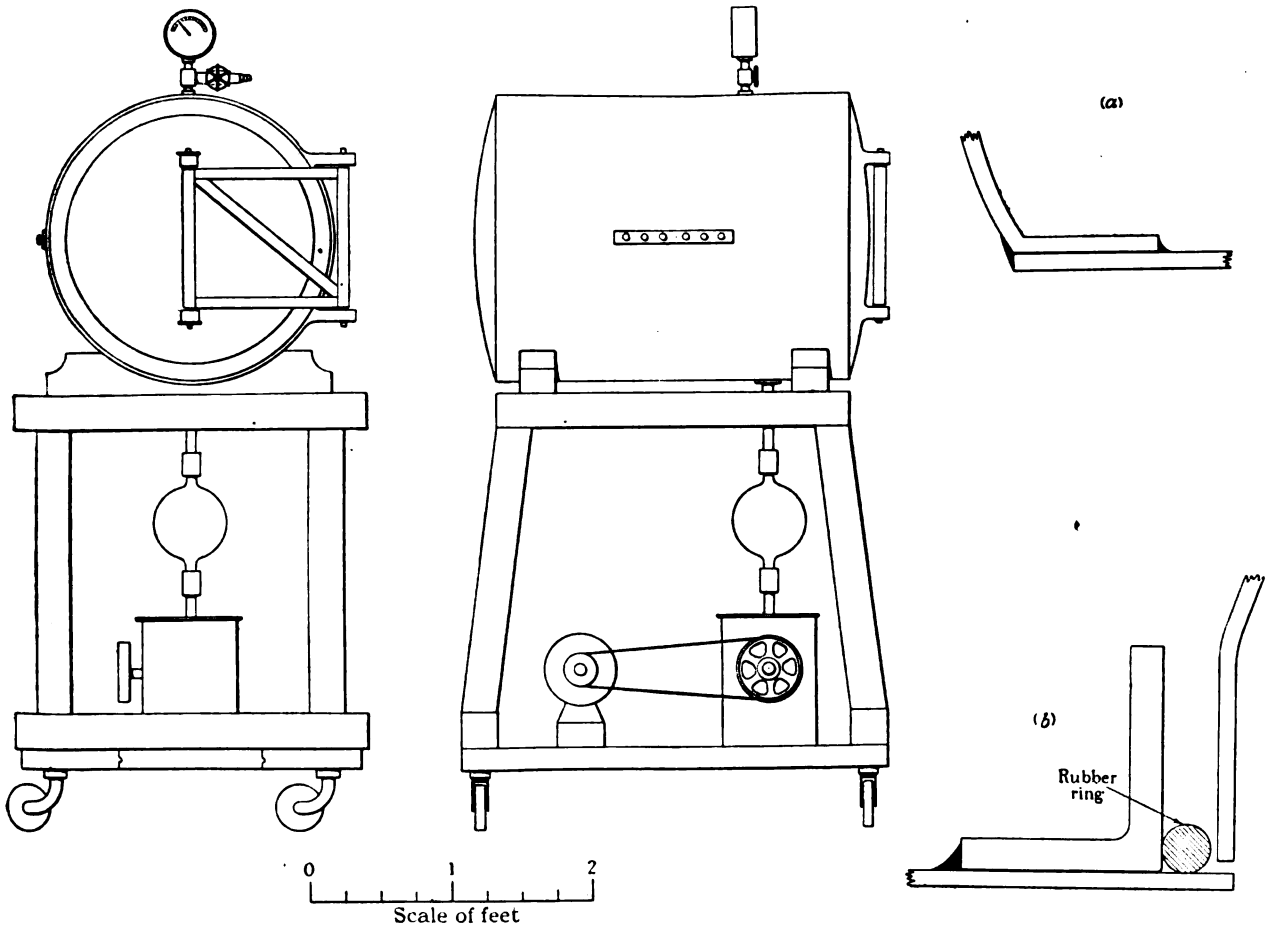

FIG. 3.—The vacuum chamber.

Details of (a) Rear-end joint; (b) Door seal.

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R_2 and X_2 (" normal "). In working condition, with a square baffle of 3 ft. side. Tested in a large room, 20 ft. × 40 ft., about 20 ft. high. No special precautions were then taken to avoid reflection from the walls on to the speaker, after preliminary tests had shown no perceptible changes of input impedance from this cause. It was found that, at low frequencies, holding a flat sheet of wood of about 2 sq. ft. within 2 or 3 ft. of the instrument would unbalance the bridge, but that the effect was not perceptible at greater distances.

R_3 and X_3 (" vacuum "). The speaker (without baffle) was inserted in a chamber maintained at more than 29 in. of vacuum, provided with air-tight entry for connections.

(4) The Measuring Equipment.

The Bridge Outfit.

This, which was a semi-permanent assembly used for practically all the bridge work of the laboratory, was a particularly convenient one. It is shown in skeleton form in Fig. 2. The bridge network itself is of a normal type, except that a reversing switch connected at points A and B can interchange the two half-primaries of the mutual inductor when it is desired to measure a negative reactance. The Wagner earth is switched in by C.

The switch box changes source and detector from the main bridge to the Campbell frequency bridge. For frequencies above the limit of this (4 000 cycles per sec.), the author relied on the calibration of the source, which had been done by Lissajous figures in a Western cathode-ray oscillograph, against an auxiliary oscillator. The switch box also contains variable-ratio screened transformers.

The output amplifier was a 3-valve one of considerable power, and is, in the author's opinion, an essential part of any bridge equipment—not so much on account of the increased loudness *per se*, which may actually be a disadvantage, but because, by bringing up previously inaudible out-of-balance signals to clear audibility, it increases the sharpness of balance some hundred-fold.

The source is a beat oscillator designed with much care. On full power it can generate about 0·2 watt with less than 1·5 per cent harmonic, and its output is within ± 5 per cent of normal between 20 and 10 000 cycles per sec. The frequency scale is approximately logarithmic, and after allowing for initial drift after starting up (by re-setting the zero) it maintains its frequency for long periods within one or two cycles.

For detection, a Campbell vibration galvanometer was used for frequencies below 150. As this would not normally go below 80 cycles per sec., the movement was loaded with two small lumps of cement, which gave the desired low range. For frequencies between 150 and 2 000 cycles per sec., telephones were used. As these were insensitive to higher frequencies, the bridge output was mixed with that from an auxiliary oscillator, and the whole rectified by an additional triode, the auxiliary oscillator being adjusted to beat with the bridge output at about 700 cycles per sec.

Considerable difficulty was at first experienced in getting a suitable vacuum chamber, as at first the quotations obtained were all for heavy apparatus of the autoclave type, at prohibitive prices.

Finally, however, Messrs. Edwards and Co. of Denmark Hill grasped the requirements, and built the chamber shown diagrammatically in Fig. 3. The chamber itself was 2 ft. diameter by 2 ft. 6 in. long, of ⅛ in. mild steel. The rear end was welded on, as also was the internal flange for the front door. The door itself rested on a soft rubber gasket, and when this was lightly greased the chamber would hold the rough vacuum required for several hours. Six terminals were fitted, with a gastight lead-in to six corresponding terminals inside.

A standard Cenco Hyvac pump, type " A," would bring the pressure down to 1 mm or so in 20 minutes, and as this was all that was needed in " height " of vacuum an ordinary dial vacuum gauge was sufficient. For convenience, the whole outfit was mounted on rubber-tyred castors.

(5) The Observations.

(a) Difficulties of Observation.

The above-described bridge equipment had given excellent consistency and accuracy on many measurements, including some rather difficult ones on the residual resistance of high reactances. The accuracy of the present measurements was not, however, of the same high order, mainly from a quite unexpected effect —the sensitivity of the loud-speaker as a microphone. The slightest disturbance, either mechanical or acoustic, caused it to impress such e.m.f.'s on the bridge that they drowned the normal note when trying to get accurate balance. Even in the vacuum chamber, where acoustic effects were non-existent, the speaker had to be supported on thick layers of sponge rubber, and no one could take a step on the wooden floor within 30 yards without causing disturbance. The " normal " tests, which were the most difficult, were made on a Sunday, with the works idle; but the day happened to be windy, and the first few tests at low frequency, using the vibration galvanometer, each took about forty minutes. Eventually a device was used which is believed to be new. There was an Ashdown rotoscope in the laboratory; this was set at nearly twice the frequency being used and the vibrating image was examined with it. This reduced the band of light to two lines of light slowly approaching and receding from one another, and made it comparatively easy to distinguish between the normal constant-frequency movement and the random ones due to disturbance.

It is a further misfortune that the interpretation of the measurements depends throughout on the differences between the three sets of readings, which naturally decreases the accuracy of the results.

These final results, however, do enable us to derive some interesting and quite definite conclusions about the speaker in the range 40–800 cycles per sec., and to make some equally interesting but more speculative inferences about higher frequencies.

(b) Preliminary Treatment of the Data.

In order to avoid accidental divergences as far as possible, the observed values were first plotted on a large scale, curves were drawn, and a series of " probable " values were derived from the curves for values

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TABLE 1.

Observed Values.

Held			Normal			Vacuum		
f	R_1	X_1	f	R_2	X_2	f	R_3	X_3
cycles/sec.	ohms	ohms	cycles/sec.	ohms	ohms	cycles/sec.	ohms	ohms
35*	2·09	0·107	39	2·34	−3·19	40	2·80	−10·60
65*	2·09	0·157	58	2·23	−1·50	62	2·22	− 3·04
103	2·11	0·040	80	2·25	−0·91	78	2·23	− 2·02
150	2·13	0·207	99	2·25	−0·63	101	2·20	− 1·39
200	2·14	0·425	121	2·28	$-0\cdot49_5$	120	2·21	− 1·04
300	2·16	0·425	139	2·25	$-0\cdot33_6$	150	2·22	− 0·70
400	2·18	0·562	161	2·29	$-0\cdot21_4$	200	2·30	$-0\cdot37_7$
500	2·22	0·690	200	2·30	$-0\cdot09_4$	298	2·27	0·00
700	2·28	0·940	221	2·27	0·00	400	2·35	$+0\cdot32_1$
1 000	2·36	1·290	300	2·26	$+0\cdot23_7$	500	2·39	$+0\cdot49_0$
			400	2·30	$+0\cdot41_2$	700	2·44	+ 0·80
			500	2·34	$+0\cdot59_0$	1 000	2·51	+ 1·22
			700	2·38	$+0\cdot85_3$	1 500	2·65	+ 1·84
			1 000	2·44	+1·22	2 000	2·75	+ 2·51
			1 500	2·58	+1·84	3 000	3·20	+ 2·82
			2 000	2·70	+2·38	4 000†	3·3	+ 3·75
			3 000	3·60	+3·08	5 000†	3·5	+ 4·07
			4 000†	3·5	+3·51	6 000†	4·0	+ 3·75
			6 000†	4·0	+1·88			

* Measurements difficult. † Resistance balance flat.

TABLE 1A.

Most Probable Values.

(Used for computation.)

f	ω	X_1	X_2	X_3	R_1	R_2	R_3
cycles/sec.	$10^2\times$	ohms	ohms	ohms	ohms	ohms	ohms
39·8	0·25	+0·055	−3·05	−11·0	$2\cdot09_0$	$2\cdot34_0$	$2\cdot80_0$
47·8	0·30	0·066	−2·05	− 5·6	$2\cdot09_0$	$2\cdot28_5$	$2\cdot45_0$
63·7	0·40	0·088	−1·30	− 2·85	$2\cdot09_5$	$2\cdot25_0$	$2\cdot26_0$
79·6	0·50	0·111	−0·92	− 2·00	$2\cdot10_0$	$2\cdot25_0$	$2\cdot22_0$
111	0·70	0·154	−0·55	− 1·18	$2\cdot11_3$	$2\cdot25_8$	$2\cdot20_5$
159	1·0	0·220	−0·24	− 0·63	$2\cdot12_7$	$2\cdot27_0$	$2\cdot22_0$
239	1·5	$0\cdot33_0$	+0·06	− 0·20	$2\cdot15_5$	$2\cdot29_0$	$2\cdot26_5$
318	2	$0\cdot46_0$	+0·27	+ 0·08	$2\cdot17_5$	$2\cdot30_8$	$2\cdot31_0$
478	3	0·66	+0·56	+ 0·45	$2\cdot20_2$	$2\cdot33_3$	$2\cdot38_5$
637	4	0·86	+0·78	+ 0·70	$2\cdot24_2$	$2\cdot36_5$	$2\cdot42_3$
796	5	1·07	$+0\cdot99_5$	+ 0·92	$2\cdot28_5$	$2\cdot40_0$	$2\cdot46_0$
1 110	7		+1·40			2·54	
1 590	10		+2·00			2·60	
2 390	15		+2·80			2·86	
3 180	20		+3·25			3·12	
4 780	30		+3·7			3·85	
6 370	40		+3·8			4·2	
7 960	50		+3·9			4·4	

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of f giving convenient values of ω. These latter were used as the basis of the following work.

It is fully realized that in adopting this procedure the author might be masking the symptoms of real effects. But a lengthy and detailed consideration, both of the actual data and of the results of analysis, failed to show correlation between the apparent divergences and any causes which the author could imagine to account for them; and it is believed that the course adopted was the best one. The original data are given *in extenso* in Table 1.

Table 1A gives the "probable" values on which further work was done, and Figs. 4 and 5 show them graphically. On these curves the original observations are shown as crosses and points in circles, and the curves themselves are the source of the "probable" values. The points in diamonds are the values found by re-computation, as will be described in Section (10). It will be seen that the reactance curves deviate little from the observed values, except (1) the small values of X_1, the "held" reactance, and (2) the values at high frequencies. The curve for X_1 at low frequencies

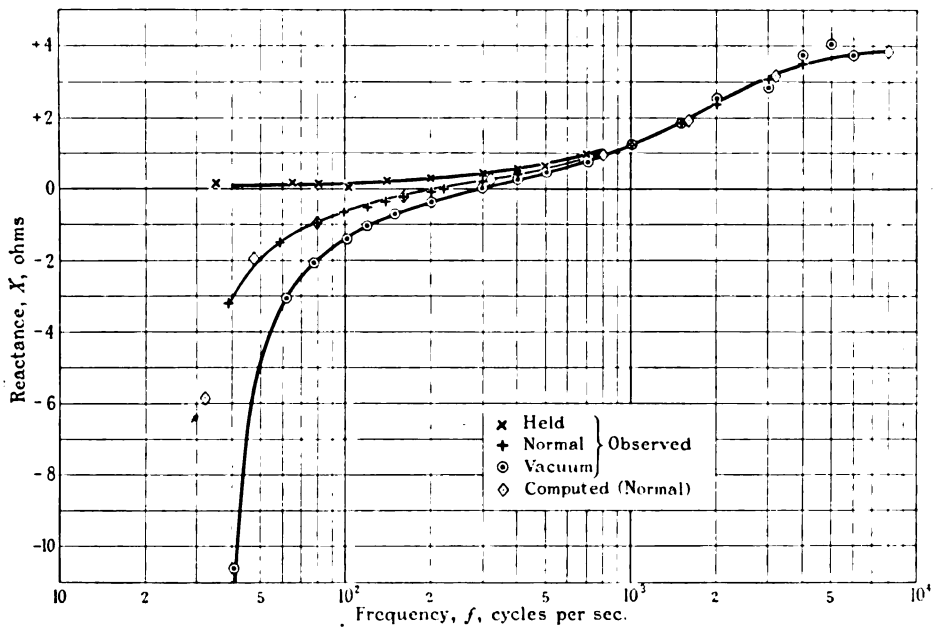

FIG. 4.—Reactance values.

All observations are plotted. The curves were drawn as the best fit, and values taken from the curves were used as the basis of computation (between 40 and 800 cycles per sec.). The points in diamonds are the result of recomputation, as explained in the text.

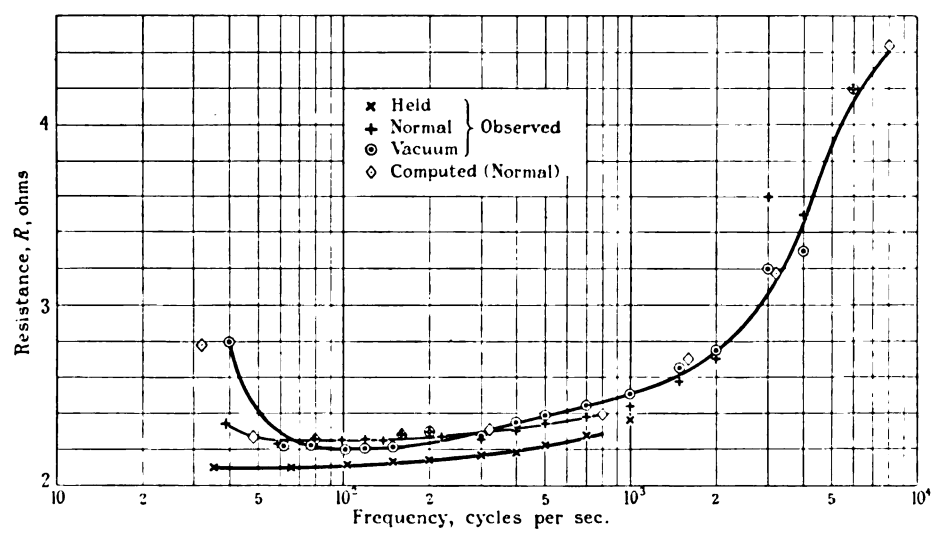

FIG. 5.—Resistance values (for remarks see Fig. 4).

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was drawn by re-plotting on metric paper X_1 against f, on the hypothesis that for such low frequencies L would be constant. L was found by the best straight line (see Fig. 6) to be 223 μH at frequencies below 400 cycles per sec., and the corresponding part of the curve for X_1 was plotted on this basis. The inductance of a sister coil in air was 126 μH.

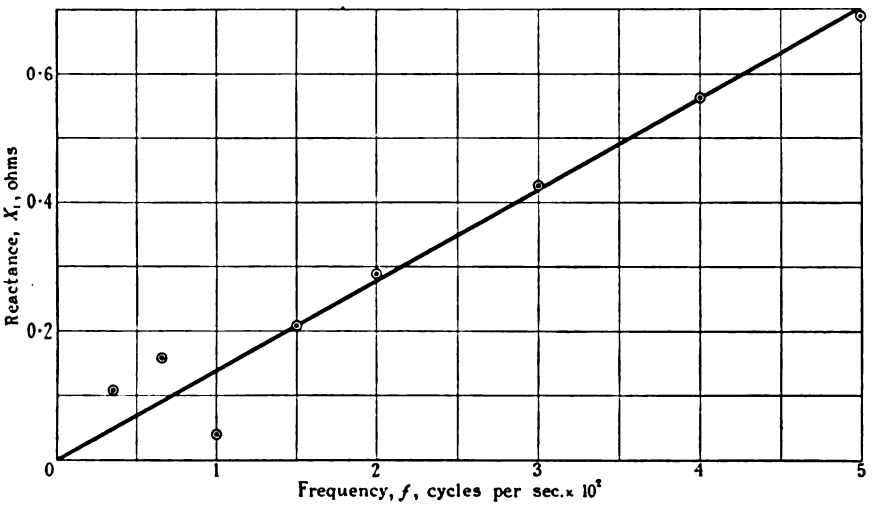

FIG. 6.—Reactance, X_1, plotted against ω at low frequencies, to find the most probable value of the coil inductance.

The differences in the reactances at high frequencies (above 1 000 cycles per sec.) appeared to be masked by errors of measurement. It was in any case to be expected that they would be small, from consideration of the fundamental equations. It was decided to plot all available values and take the best curve, ignoring the true differences.

In the resistance values the same irregularities occur at high frequencies; the resistance balance was, in fact, very flat at over 3 000 cycles per sec. The resistances at low frequencies do not seem to show large irregularities.

On consideration of the data, it was decided to compute for a series of pulsatances up to $\omega = 5\,000$, and then to investigate the possibility of extrapolating for higher frequencies.

It soon became evident that wave transmission down the cone became the dominating factor in the mechanical impedances at quite low frequencies: anticipating the results, symptoms of it were obvious at 500 cycles per sec. It is therefore essential to consider such wave transmission and its results.

(c) Wave Transmission.

The complex cone system may be resolved into the following:—

(i) A concentrated mass and compliance at the input end, corresponding to the coil and former and its spring retaining washer, together with a resistance due to any internal friction or power dissipation in bending the washer, and (in the acoustic or " normal " case) air friction in the gap.

(ii) A " transmission line."

(iii) A " terminal impedance " corresponding to the flexible surrounding material between cone and baffle.

The electrical equivalent circuit is shown in Fig. 7, in which C_1 is the " compliance " of the central washer, and R_1 any viscosity therein; L_1 the mass of coil and

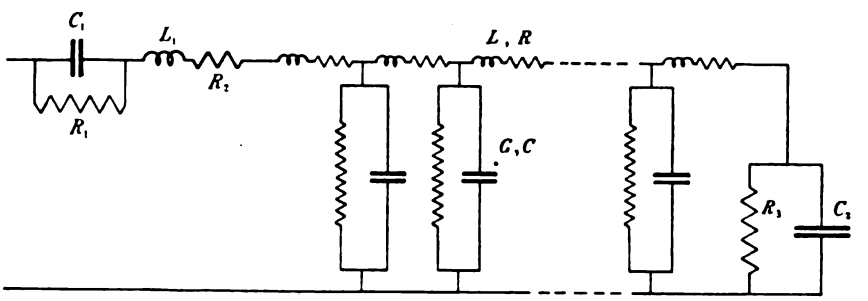

FIG. 7.—The electrical line equivalent to the cone and its suspension.

C_1 and R_1 = elastance and viscosity of centring washer.
L_1 = mass of coil and former.
R_2 = air friction in gap.
L, R, C, G = line components (L = mass of cone and acoustic mass; R = acoustic resistance; C = elastance of cone itself; G = viscosity in C; C_2, R_3 = elastance and viscosity of " surround ").
The mass of the surround is neglected, but would appear as an inductance in series with the $C_2 - R_3$ load.

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former (and part of the mass of C_1), and R_2 the effect of air friction in the gap; and C_2 and R_3 corresponding equivalents for the "surround." Strictly, part of the mass of the surround should be shown as an inductance at the end of the line, but it is neglected for the present. L, R, G and C are the four parameters for the "line" itself.

These last cause great difficulty, for they are not constant along the line. L represents the mass, and this obviously increases outwards. R is only present in the acoustic case, being representative of actual radiated power. In the acoustic case there is also an addition to L, and we know little of the distribution over the surface of either of these. C, the compliance, is a function of the shape of the cone, and will therefore itself be affected by any change of shape due to wave transmission. G represents dissipation within the cone material.

We therefore cannot expect this "line" to behave in detail like a normal one. But in the author's opinion we are justified in certain assumptions:—

(i) It seems almost certain that at very high frequencies the input impedance will tend to constancy. In the case of a normal line it tends towards $R = \text{const.}$, $X = 0$; but in other types of calculable line X also tends towards a constant finite value.

(ii) At very low frequencies we may certainly assume "rigid" motion, in which case the input impedance will simply consist of the total R plus the reactance of the total mass.

(iii) A rather more speculative assumption is that the first divergence from (ii) as frequency rises will be, to a first approximation, of the same type as that for a normal line.

Taking the vacuum case, we may put:—

$$R = 0$$

total $L = m =$ total mass
total $G = g =$ total viscosity in the cone material
total $C = c =$ total compliance

and then, for the input impedance Z of a simple line, we have

$$Z' = Z_0 \tanh \alpha \quad . \quad . \quad . \quad . \quad . \quad (15)$$

where
$$Z_0 = \sqrt{\left[\frac{j\omega m}{g + j\omega c}\right]} \quad . \quad . \quad . \quad . \quad (16)$$

and
$$\alpha = \sqrt{[j\omega m(g + j\omega c)]} \quad . \quad . \quad . \quad . \quad (17)$$

When α is so small that $\tanh \alpha = \alpha$,

$$Z' = j\omega m \quad . \quad . \quad . \quad . \quad . \quad . \quad (18)$$

This is condition (ii) above.

For rather larger values of α, we have

$$\tanh \alpha = \left(1 - \frac{\alpha^2}{3}\right)\alpha \quad . \quad . \quad . \quad . \quad (19)$$

and it is easily proved that, if $Z' = R' + j\omega m'$,

$$R' = \frac{\omega^2 m^2}{3} g \quad . \quad . \quad . \quad . \quad . \quad (20)$$

$$m' = m + \frac{m^2 c}{3}\omega^2 \quad . \quad . \quad . \quad . \quad (21)$$

It will be seen later that these relations do hold approximately.

It might be noted that equation (15) is for a line short-circuited at the far end. For any good loudspeaker, the impedance of the outer surround washer is so low that this is substantially the case.

(6) The Electrical and Mechanical Impedances.

(a) *The Inductance and Resistance of the Coil.*

Considering first the "fixed" values, it is interesting to investigate the frequency-change of L and R. If the eddy-current path in the iron around the gap were a definite one, we should have, putting L_4 and R_4 for the simple inductance and resistance of the coil itself, L_5 and R_5 for those of the eddy-current path, and M for the mutual inductance between them,

$$R_c = R_4 + \gamma^2 R_5 \quad . \quad . \quad . \quad . \quad . \quad (22)$$

$$L = L_4 - \gamma^2 L_5 \quad . \quad . \quad . \quad . \quad . \quad (23)$$

where
$$\gamma^2 = \frac{\omega^2 M^2}{R_5^2 + \omega^2 L_5^2} \quad . \quad . \quad . \quad . \quad . \quad (24)$$

By plotting $1/(R_c - R_4)$ or $1/(L_4 - L)$ against $1/\omega^2$, we should get straight lines, from which, it is found, we can derive $\delta = R_5/L_5$, and μ, the coefficient of coupling between L_4 and L_5. Further separation cannot be effected, i.e. *any* secondary of the same coupling coefficient and time-constant would give the same frequency characteristic of apparent inductance and resistance. By computing the inductance and resistance of various sections of the pole-pieces, we can find what radial thickness gives the observed order of value of δ.

In practice, the penetration, and hence the impedance, of the equivalent secondary, is itself a function of frequency, but a rough computation was carried out, and it was found that at the higher frequencies $\mu \doteqdot 80$ per cent and $\delta \doteqdot 2{\cdot}3 \times 10^4$. Utilizing the latter figure, the apparent penetration into the iron was of the order of $0{\cdot}4$ mm.

(b) *The Mass and Stiffness at Low Frequencies, and the Force Factor.*

Returning now to equation (9), we have

$$Z = R + jX = R_c + R_r + j(X_c + X_r)$$

But the "held" measurements supply values for R_c and X_c, so that we get, by subtraction, R_r and X_r. From these, by well-known formulæ for the equivalence of series and parallel circuits, we can obtain R_p and X_p and hence, by equations (10) to (12), R_d and X_d.

Perhaps the most suitable series-parallel conversion formula is found as follows.

Let
$$k = \frac{R_r}{X_r} \text{ or } \frac{X_r}{R_r} \quad . \quad . \quad . \quad . \quad (25)$$

$$q = k + \frac{1}{k} \quad . \quad . \quad . \quad . \quad . \quad . \quad (26)$$

Then
$$R_p = qX_r \quad . \quad . \quad . \quad . \quad . \quad . \quad (27)$$

$$X_p = qR_r \quad . \quad . \quad . \quad . \quad . \quad . \quad (28)$$

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Lines (2) to (8) of Table 2 carry out this computation for the "vacuum" measurements. Lines (2) and (3) give X_r and R_r by subtraction of the probable values in Table 1A. Lines (7) and (8) give R_p and X_p, and line (9) gives

$$C_p(\text{farads}) = Y_p/\omega = 1/(\omega X_p) \quad . \quad . \quad (29)$$

Now, by equation (11), we have $Y_p = X_d/D$,

or
$$C_p = \frac{m + \frac{X_a}{\omega} - \frac{s}{\omega^2}}{D} \quad . \quad . \quad . \quad . \quad (30)$$

and since in this case X_a is excluded, as we are working *in vacuo*,

$$C_p = \frac{m}{D} - \frac{s}{D} \cdot \frac{1}{\omega^2} \quad . \quad . \quad . \quad . \quad (31)$$

Hence, by plotting C_p against $1/\omega^2$ we should be able to derive m/D and s/D, in so far as they are constants. It must be remembered that above the frequency at

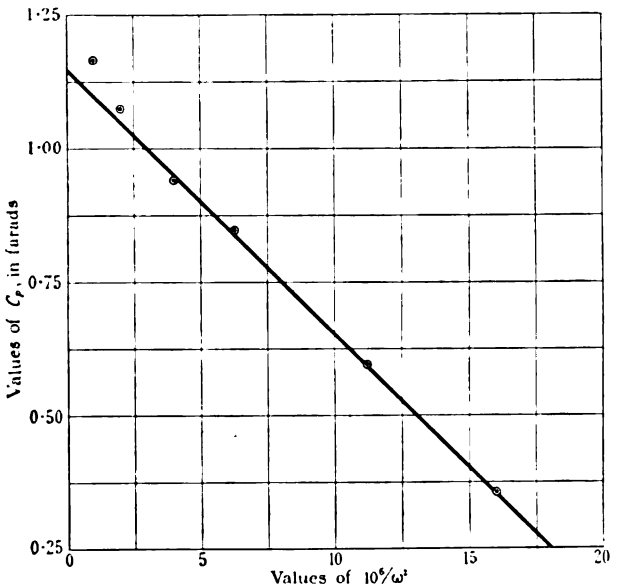

FIG. 8.—C_p plotted against $1/\omega^2$ at low frequencies, from which A and s are derived (m being known).

which wave transmission is first perceptible, m' will at first increase [see equation (21)]. But the lowest frequencies, at any rate, should show a straight line. The results are plotted in Fig. 8. The best straight line was drawn among points for low frequencies, as shown. This gave

$$m/D = 1{\cdot}15 \times 10^{-3} \quad . \quad . \quad . \quad . \quad (32)$$

$$s/(10^6 D) = 0{\cdot}05 \times 10^{-3} \quad . \quad . \quad . \quad . \quad (33)$$

Now the *static* value of m, i.e. the actual mass of the diaphragm, etc., was known to be as follows:—

Wire of coil ..	3 grammes	
Former	1 gramme	
Part of washer ..	1 ,,	
	—	
	5 grammes .	(34)
Cone ..	14 ,, .	(35)
Part of surround	2 ,, .	(36)
	—	
	21 grammes .	(37)

Hence $D = 18{\cdot}3 \times 10^3 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (38)$

$$A = \sqrt{(10^9 D)} = 4{\cdot}28 \times 10^6 \text{ C.G.S. units} \quad (39)$$

The coil consisted of 61 turns of just over 2·5 cm diameter, giving

$$l = 492 \text{ cm} \quad . \quad . \quad . \quad . \quad . \quad . \quad (40)$$

whence $B = A/l = 8\,700$ gauss . . . (41)

also, since [from equation (33)] $s = 50\,D$, we have

$$s = 0{\cdot}915 \times 10^6 \text{ dynes per cm} \quad . \quad . \quad . \quad (42)$$

Both B and s agree quite well with static measurements on sister instruments, which gave values of B varying between 8×10^3 and 10×10^3, and s from $0{\cdot}6 \times 10^6$ to 1×10^6.

(c) The Input Impedance of the Cone as a "Line."

Using the values of D and s just obtained, we shall now derive (from the measured values of C_p) the apparent value of m, which we shall call m', at varying frequencies. From equation (31) we have

$$m' = DC_p + \frac{s}{\omega^2} \quad . \quad . \quad . \quad . \quad . \quad (43)$$

Lines (10) to (12) of Table 2 show this computation. Line (13) shows R_d ($= DG_p = D/R_p$), from equations (12) and (14).

For possible extrapolation, it is useful to know the input reactance of the line. This would be $\omega m'$, but for the fact that 5 grammes out of m' is due to the concentrated mass at the input end of the line. Lines (14) and (15) of Table 2 therefore show $X_d' = \omega(m' - 5)$.

R_d, X_d', and $(m' - 5)$ are plotted against ω in Figs. 9 and 10, to indicate the general nature of their variation. R_d appears parabolic at low frequencies, and, although the random errors in m' are large, there are indications that this also follows a parabolic law up to $\omega \doteqdot 10^3$. If we plot R_d and m' against ω^2, as shown in Fig. 11, we find to a first approximation the line constants of the cone (see Fig. 7). The actual figures obtained are

$$g = 24 \times 10^6$$
$$c = 0{\cdot}35 \times 10^{-8}$$

or,

stiffness of cone ($= 1/c$) $= 2{\cdot}8 \times 10^8$ dynes per cm . (44)

which is a very plausible value. The author has no information which might give a check on the plausibility of the value of g, which is the resistance corresponding to power dissipated in the actual bending of the cone material.

(7) THE ACOUSTIC IMPEDANCE.

(a) The Acoustic Mass and Effective Cone Surface.

We are now in a position to investigate the acoustic case, and in Table 3 will be found a computation similar to that of Table 2, but utilizing the data for "normal" instead of "vacuum" operation. Lines (13), (14) and (15) give the results, which have the suffix 2 to indicate that they are derived from the original measurements also given that suffix.

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TABLE 2.—*Diaphragm Impedance.*

(1)	$\omega = 10^3 \times$	0·25	0·3	0·4	0·5	0·7	1·0	1·5	2	3	4	5
(2)	$X_r =$	−11·06	− 5·67	− 2·94	− 2·11	− 1·33	−0·85	−0·55	−0·38	−0·21	−0·16	−0·15
(3)	$R_r =$	0·71	0·36	$0·16_5$	$0·12_0$	$0·09_2$	$0·09_3$	0·11	$0·13_5$	$0·18_3$	$0·18_1$	$0·17_5$
(4)	$k =$	−15·8	−15·5	$-17·8_3$	−17·6	−14·3	−9·14	−5·0	−2·82	−1·15	$-0·88_5$	$-0·86_0$
(5)	$1/k =$	− 0·06	− 0·06	− 0·06	− 0·06	− 0·07	−0·11	−0·2	−0·36	−0·87	−1·13	−1·17
(6)	$q =$	−15·86	−15·56	−17·89	−17·66	−14·37	−9·25	−5·2	−3·18	−2·02	$-2·01_5$	−2·03
(7)	$R_p =$	175·2	88·1	52·6	38·2	$19·2_4$	7·86	2·86	1·21	0·424	0·321	0·305
(8)	$X_p =$	−11·25	− 5·60	− 2·95	− 2·12	− 1·33	−0·86	$-0·57_2$	$-0·43_0$	$-0·37_0$	$-0·36_4$	$-0·35_5$
(9)	$C_p = 10^{-3} \times$	0·356	0·595	0·848	0·943	1·075	1·164	1·167	1·163	0·900	0·688	0·564
(10)	$DC_p =$	6·50	$10·9_0$	$15·5_3$	$17·1_5$	$19·6_5$	$21·3_5$	$21·4_0$	$21·3_5$	$16·4_5$	$12·5_7$	$10·3_3$
(11)	$s/\omega^2 =$	$14·6_5$	$10·1_5$	$5·7_1$	3·65	1·86	0·92	0·41	0·23	0·10	0·06	0·04
(12)	$m' =$	$21·1_5$	$21·0_5$	$21·2_4$	$20·8_0$	$21·5_1$	$22·2_7$	$21·8_1$	$21·5_8$	$16·5_5$	$12·6_3$	$10·3_7$
(13)	$R_d = 10^3 \times$	$0·104_2$	0·208	0·348	0·480	0·95	2·33	6·40	$15·1_5$	43·3	57·1	60·0
(14)	$m' - 5 =$	$16·1_5$	$16·0_5$	$16·2_4$	$15·8_0$	$16·5_1$	$17·2_7$	$16·8_1$	$16·5_8$	$11·5_5$	$7·6_3$	$5·3_7$
(15)	$X'_d = 10^3 \times$	4·04	$4·8_2$	$6·5_0$	$7·9_0$	$11·5_5$	$17·2_7$	$25·2_0$	$33·1_6$	$34·6_5$	$28·5_2$	$26·8_5$

TABLE 3.—*Acoustic Reactance.*

(1)	$\omega = 10^3 \times$	0·25	0·3	0·4	0·5	0·7	1·0	1·5	2	3	4	5
(2)	$X_{r_2} =$	− 3·10	− 2·12	−1·39	−1·03	$-0·70_4$	−0·46	−0·29	−0·19	−0·10	−0·08	$-0·07_5$
(3)	$R_{r_2} =$	0·25	0·195	0·155	0·150	0·145	0·143	0·135	0·133	0·131	0·125	0·115
(4)	$k =$	−12·40	−10·88	−8·98	−6·88	−4·85	−3·22	−2·15	−1·43	−0·76	−0·64	−0·65
(5)	$1/k =$	− 0·08	− 0·09	−0·11	−0·15	−0·21	−0·31	−0·47	−0·70	−1·31	−1·56	−1·53
(6)	$q =$	−12·48	−10·97	−9·09	−7·03	−5·06	−3·53	−2·62	−2·13	−2·07	−2·20	−2·18
(7)	$R_{p_2} =$	38·7	23·3	$12·6_5$	7·24	3·56	1·62	0·76	$0·40_5$	$0·20_7$	$0·17_6$	$0·16_4$
(8)	$X_{p_2} =$	− 3·12	− 2·14	−1·41	$-1·05_5$	$-0·73_4$	$-·50_5$	$-0·35_3$	$-0·28_3$	$-0·27_1$	$-0·27_5$	$-0·25_1$
(9)	$C_{p_2} = 10^{-3} \times$	$1·28_3$	$1·56_0$	$1·77_5$	$1·89_6$	$1·95_0$	1·98	1·89	1·77	1·23	0·91	$0·79_7$
(10)	$DC_{p_2} =$	23·5	$28·5_5$	32·5	34·6	35·6	36·3	34·6	32·4	22·5	16·6	14·6
(11)	$s/\omega^2 =$	$14·6_5$	$10·1_5$	$5·7_0$	$3·6_5$	$1·8_6$	$0·9_0$	0·4	0·2	0·1	0·06	0·04
(12)	$m'_2 =$	$38·1_5$	$38·7_0$	$38·2_0$	$38·2_5$	$37·4_6$	37·2	35·0	32·6	22·6	$16·6_6$	$14·6_4$
(13)	$R_{d_2} = 10^3 \times$	0·47	$0·78_5$	1·44	2·53	5·15	11·3	23·1	45·2	88	104	112
(14)	$m'_2 - 5 =$	$33·1_5$	$33·7_0$	$33·2_0$	$33·2_5$	$32·4_6$	32·2	30·0	27·6	17·6	$11·6_6$	$9·6_4$
(15)	$X'_{d_2} = 10^3 \times$	8·29	10·1	13·3	16·6	22·7	32·2	45	55·2	52·8	46·8	48
(16)	$X'_{d_3} = 10^3 \times$	4·04	$4·8_2$	6·5	7·9	$11·5_5$	17·3	25·0	33·2	34·6	28·1	26·8
(17)	$X_a = 10^3 \times$	4·25	$5·2_8$	6·8	8·7	$11·1_5$	14·9	20·0	22·0	18·2	18·7	21·2
(18)	$X_a/S =$	3·5	4·2	5·6	7·0	9·6	13	18	23	29	$29·0_5$	25
(19)	$X'_a = 10^3 \times$	4·37	5·25	7·0	$8·7_5$	12·0	16·2	22·5	28·7	36·3	37	31·4

TABLE 4.—*Acoustic Resistance.*

(1)	$\omega = 10^3 \times$	0·25	0·3	0·4	0·5	0·7	1·0	1·5	2	3	4	5
(2)	$R_{d_2} = 10^3 \times$	0·47	$0·78_5$	1·44	2·53	5·15	11·3	23·1	45·2	88	104	112
(3)	$R_{d_3} = 10^3 \times$	$0·10_4$	$0·20_8$	0·35	0·48	0·95	2·30	6·4	15·2	43·3	57	60
(4)	$R_a = 10^3 \times$	$0·36_6$	$0·57_7$	1·09	2·05	4·20	9·00	16·7	30·0	44·7	47	52
(5)	$R_a/S =$	0·19	0·28	0·5	0·8	1·5	2·9	6·0	10	21	33	42
(6)	$R'_a = 10^3 \times$	$0·23_6$	0·35	$0·62_5$	1·00	1·88	3·63	7·5	12·5	26·2	41·3	52·5
(7)	$R_a/R'_a =$	1·55	1·65	1·74	2·05	2·23	2·5	2·23	2·4	1·7	1·14	1

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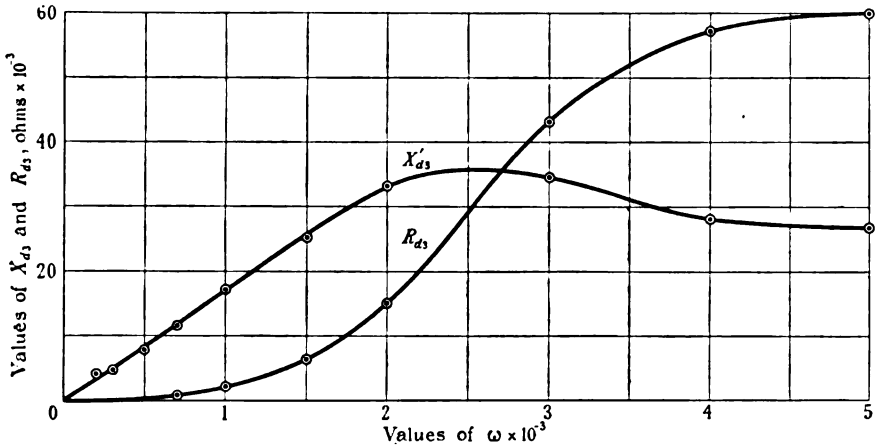

FIG. 9.—The mechanical resistance and reactance of the cone, derived from the vacuum measurements.

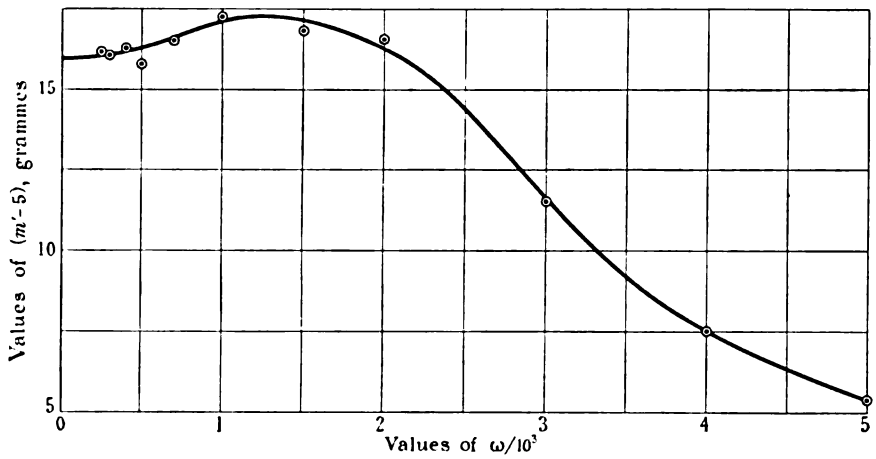

FIG. 10.—The effective input mass of the cone, derived from X_{d3} (Fig. 9).

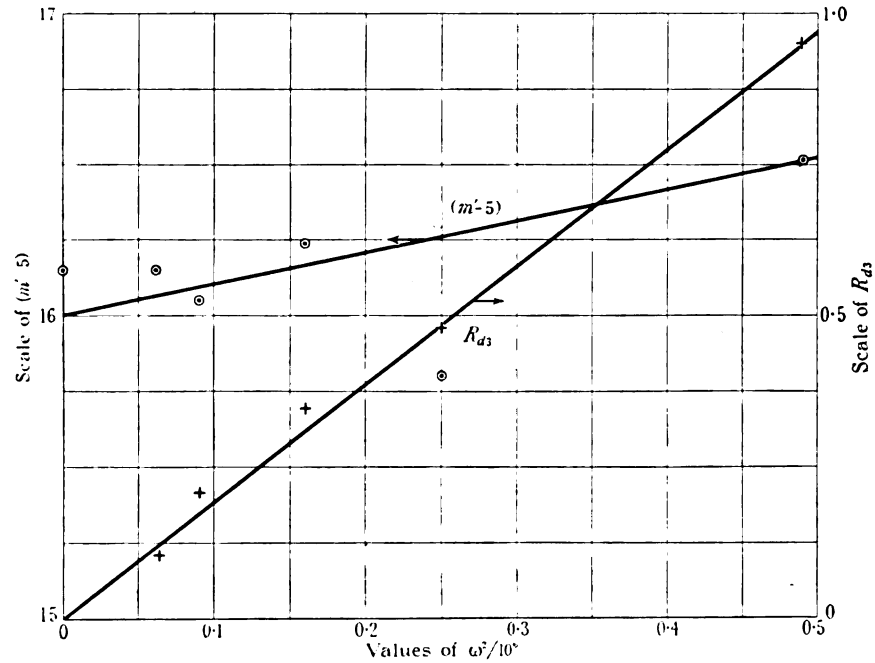

FIG. 11.—R_{d3} and m' plotted against ω^2 for low frequencies, giving a first approximation to the stiffness and viscosity-resistance of the cone.

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Fig. 12 shows curves of X'_{d_2} and R_{d_2} against ω, with X'_{d_3} and R_{d_3} (from Fig. 10) shown dotted for comparison, while Fig. 13 shows $m'_2 - 5$, with $m'_3 - 5$ dotted.

First, the values for $m'_2 - 5$ for low frequencies will be considered. The curve of Fig. 1, for X_d/S, is linear for low frequencies; in fact, if m_a is the " acoustic mass," we have

$$m_a = 0{\cdot}0138\, S \text{ (grammes)} \quad . \quad . \quad . \quad (45)$$

This is for both sides of the cone, so that the effective area of one side is 625 cm^2, corresponding to a radius of 14·1 cm. Now the projected radius of the cone itself was 12·5 cm, and the surround was 2 cm wide, so that, in this case at any rate, the effective area at low frequencies is to a close approximation the projected area of the cone plus three-quarters of the surround.

No attempt was made, on the lines of that in Section (6 c), to derive from R_{d_2} and X'_{d_2} the " line " characteristics of the cone when acoustically loaded.

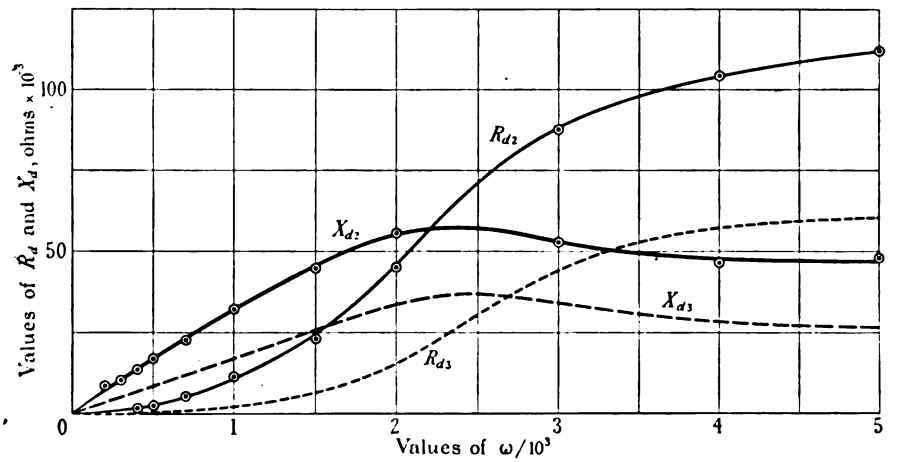

FIG. 12.—The input resistance and reactance of the cone with normal acoustic load. The pure mechanical values from Fig. 9 are shown dotted for comparison.

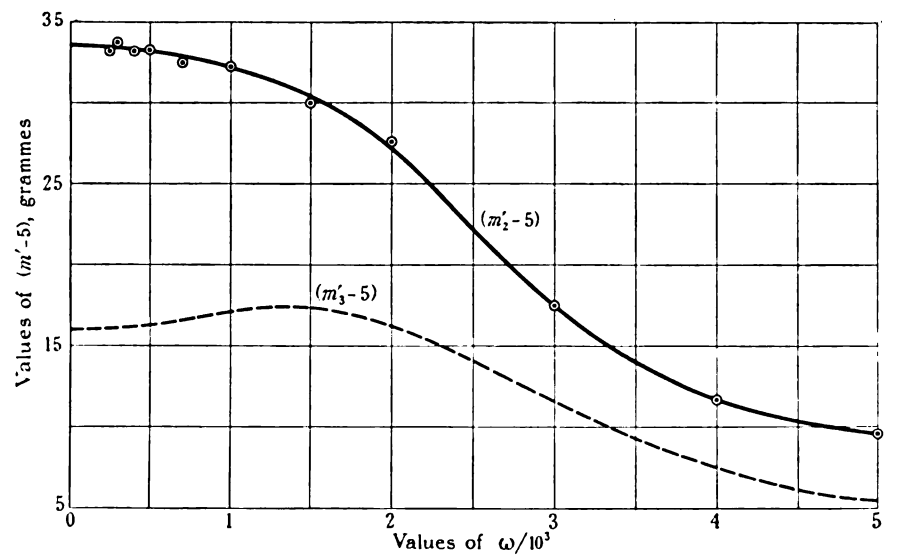

FIG. 13.—The effective input mass, derived from X_{d_2} of Fig. 12, with the " vacuum " value from Fig. 10 shown dotted.

Now the average value of $m_2 - 5$ for $\omega > 500$ is 33·26, and that of $m_s - 5$ is 16·05. Hence

$$m_a = 33{\cdot}26 - 16{\cdot}05 = 17{\cdot}21. \quad . \quad . \quad (46)$$

From this and equation (45),

$$S = 1\,250 \quad . \quad . \quad . \quad . \quad . \quad . \quad (47)$$

Such an attempt would be useless at present, as the acoustic loading is itself a function of frequency; moreover, its distribution over the cone appears to be unknown.

It also follows from the above that the computed differences between X'_{d_3} and X'_{d_2}, and R_{d_3} and R_{d_2}, are are only the same things as the acoustic reactance and resistance of the cone for frequencies so low that the

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cone behaves as a rigid body. Referring to Fig. 7, the acoustic resistance and reactance represent additions to R and L along the line, whereas the above differences represent changes in the apparent input impedance—a quite different thing.

(*b*) *The Acoustic Resistance.*

Notwithstanding the remarks in the last paragraph, it remains true that the power passed into the " line " is strictly expressed if we multiply the input resistance by the square of the current measured at the input (or, in the mechanical case, by the square of the R.M.S. velocity), and in the acoustic case the power given to the air is similarly derived from R_a.

Now the curves in Fig. 1 give the values of R_a and X_a for a given area and frequency, *for a rigid disc*. In line (18) of Table 3 are given values of X_a/S from Fig. 1, and from these we derive, in line (19), the " rigid " values of X_a', the acoustic reactance of our cone if rigid movement continued up to 800 cycles per sec. In the first five lines of Table 4 we compute similarly R_a and R_a', the latter being the theoretical acoustic resistance of a rigid disc.

At once we are on the track of a proximate cause for the unexpectedly good performance in practice of these loud-speakers (i.e. the moving-coil type in general) at is directly producing radiation: the component OC is producing skin friction.

At the same time, these " sliding " components of motion, of both coil and cone, must be producing *some* radiation, since they are doing work on the air; for a

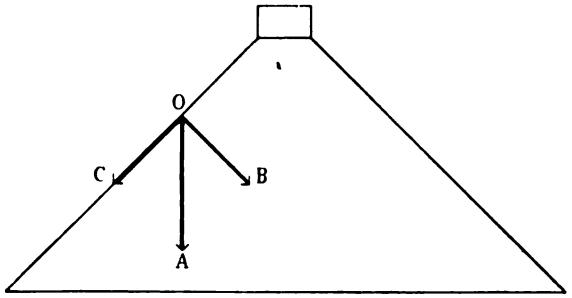

FIG. 14.—Normal and tangential components of cone motion (see text).

body can only do work by air friction if it carries some air with it, and such movements of air themselves set up radiation.

There is also the question whether the total theoretical radiation from a cone is really substantially identical with that of a disc of equal radius: to the best of the

TABLE 5.

Extrapolated Data.

(1)	$f \div$	32	48	80	160	320	800	1 600	3 200	8 000
(2)	$\omega = 10^3 \times$	0·2	0·3	0·5	1	2	5	10	20	50
(3)	$R_{d_2} = 10^3 \times$	0·36	0·78	2·53	11·3	45	112	121	124	125
(4)	$X'_{d_2} = 10^3 \times$	6·65	10·1	16·6	32·2	55·2	46	43	42	42
(5)	$R_c \mp$	2·09	2·09	2·10	2·13	2·18	2·28	2·60	3·12	4·4
(6)	$X_c =$	0·04	0·07	0·11	0·22	0·46	1·07	2·00	3·25	3·9
(7)	$R_a = 10^3 \times$	0·26	0·58	2·05	9·0	30	51	54	56	57
(8)	$R_u = 10^3 \times$	0·18	0·40	1·44	6·3	21	36	38	39	40

high frequencies. We saw in Section (6 *a*) that the electrical impedance of the coil itself falls at high frequencies; we found (see Table 2, line 14) that at 800 cycles per sec. the apparent mass of the cone from the input point of view had fallen from 16 to 5·37 grammes, and in Table 3 we saw that the acoustic reactance falls to far below its value for a rigid disc: in fact, all the " masses " giving useless reactance fall as frequency goes up. But from Table 4 we see that the useful component of impedance—the acoustic resistance—is in excess of its theoretical value.

It must not be assumed, however, that all the power expended in R_a is actually used in radiation. Some of it is certainly dissipated in " skin friction " of air about the coil in the gap. Further, it must be remembered that actually the cone is not moving normal to its own surface. This means (see Fig. 14) that of the motion of any point O of the cone, which is in the direction of OA, only the component OB, normal to the surface, author's knowledge no analysis of the problem has yet been published.

The above points are important, as any estimate of radiated power or efficiency depends on them. Obviously, the total dissipation R_a will be used in computing the motion of the diaphragm; but only the useful radiation resistance, which we may call R_u, will be employed in computing the output power.

The point will be further discussed later.

(8) COLLECTION OF COMPUTED DATA, AND EXTRAPOLATION.

The next part of the investigation is to utilize the data which have been found to re-compute the impedance of the speaker as a check, and to find the power output, for an extended frequency range if possible. It is fully realized that extrapolation to higher frequencies is of uncertain accuracy, so that this part of the paper must be regarded as speculative. None the less, it is the

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author's view that the extrapolation can be done on a very reasonable basis, and that the results are probably correct within, say, a ratio of 2 to 1 up or down—

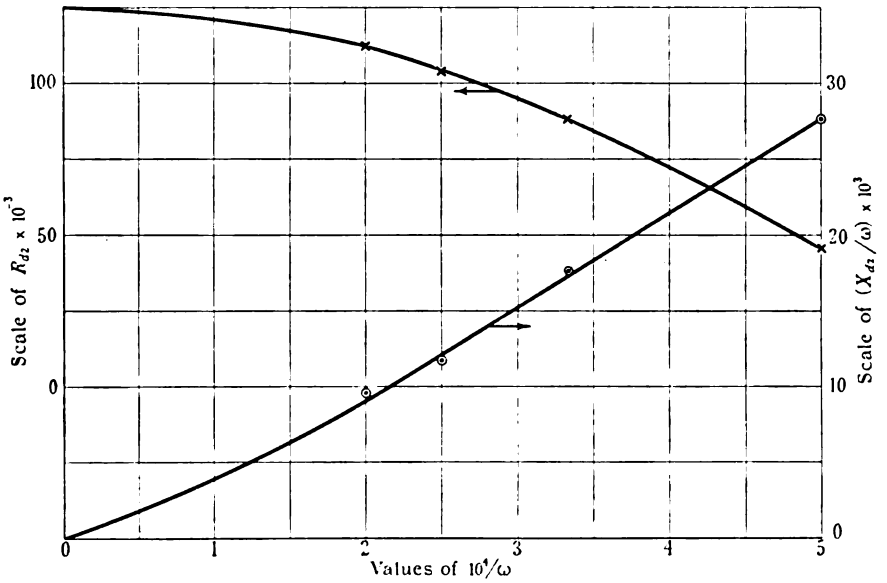

FIG. 15.—R_{d_2} and X_{d_2} plotted against inverse frequency, for extrapolation as $\omega \to \infty$.

which, we must admit with shame, is quite good accuracy in loud-speaker work.

Sketching the computation in summarized form, we have to use R_{d_2} and X_{d_2} to find the diaphragm impedance, convert to electric impedance, and add that of the coil. We then assume an input voltage and find the current. This is converted to mechanical force by the force factor A, and from the force and the diaphragm impedance we find the velocity. The square of the latter multiplied by the useful acoustic resistance gives the output power.

We can then find the true power efficiency, as the ratio of output to input power. But this is not the best criterion of performance. As is now well known, true efficiency is not so important as the relative output for a fixed input voltage or current. Now at any point in space around the loud-speaker we shall have (apart from standing-wave effects) air pressure proportional to the square root of output power, and therefore the best

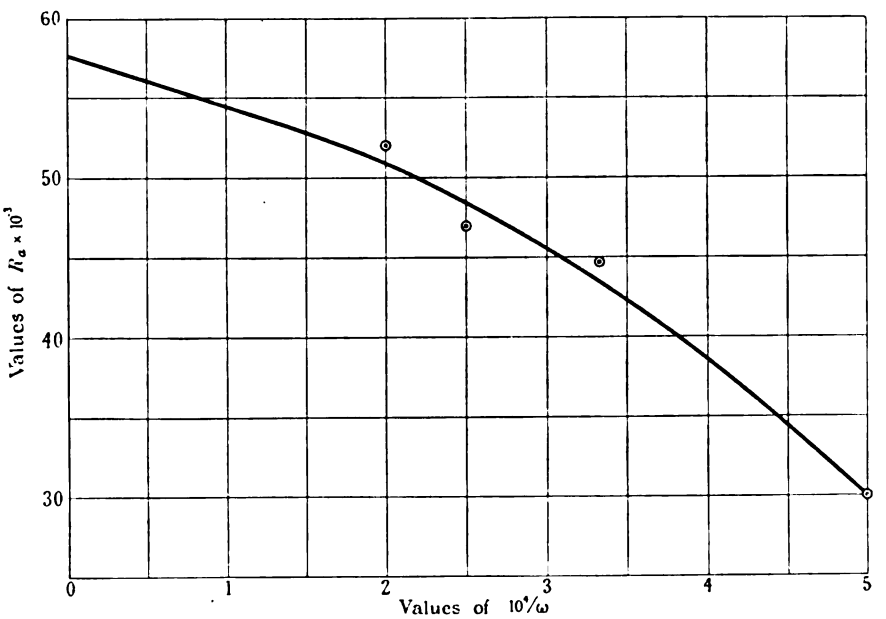

FIG. 16.—R_a plotted against inverse frequency, for extrapolation as $\omega \to \infty$.

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criterion of performance against frequency is the " output ratio "

$$\eta = \frac{\text{square root of output power}}{\text{input voltage}} \quad . \quad . \quad (48)$$

(If the instrument is to be used under " constant-current " conditions the denominator will be " input current.") This ratio, plotted against frequency, corresponds to the usual frequency/response curve of an amplifier or the like.

Let us first collect our data, and discuss the extrapolation of such of them as vary with frequency. They are set out in Table 5.

R_{d_2} *and* X'_{d_2}.—Curves of these up to $\omega = 5 \times 10^3$ are given in Fig. 12. We must bear in mind that they are the components of the input impedance of a " line." It is a general and inherent property of such that at high frequencies, so high that the line is " long," i.e. contains several waves complete, they tend towards constancy. The rise of the X-component to an intermediate maximum is also qualitatively typical.

It is obvious that these two components are merely hindrances to the output of the instrument (except only as to the useful part of R_{d_2}). It is therefore expedient to let the extrapolation be on the high side. But it will also appear that at high frequencies the reflected electrical impedance is in any case so small a fraction of the total impedance that no important error in the resultant current will result from reasonable errors in R_{d_2} and X_{d_2}. The extrapolation for R_{d_2} was done by plotting it against $1/\omega$, and extending the curve to zero as shown in Fig. 15. For X'_{d_2}, the corresponding mass was plotted against $1/\omega$, and the curve extended to pass through the origin, which is the obvious condition for constant X as $\omega \to \infty$. This is also shown on Fig. 15, and the resultant values are shown in lines (2) and (3) of Table 5.

R_c *and* X_c.—For these, the values obtained in the original measurements were smoothed, as shown in Table 1A. They had not to be widely extrapolated, as the measurements extended to $\omega = 3 \times 10^4$.

R_u.—Considerable thought was expended on this. Firstly, R_a was extrapolated in the same manner as R_{d_2} above (see Fig. 16). This was considered reasonable, for the theoretical curve of Fig. 1 shows constancy as $\omega \to \infty$, and the experimental curve also does so. As a correction for waste, it was eventually decided to multiply R_a as thus found by 0·7—approximately using the normal component only. It is realized that this is really arbitrary. It assumes that neglect of radiation from the tangential component offsets neglect of loss due to frictional loss round the coil. It will, of course, be realized that the magnitude of this " waste factor " does not affect the shape of the response curve which is our object, if the waste factor is independent of frequency. There seems no reason why it should depend on frequency, but it may do, and the final response curve is uncertain to this extent. The only possible check would seem to be a very thorough set of direct output measurements.

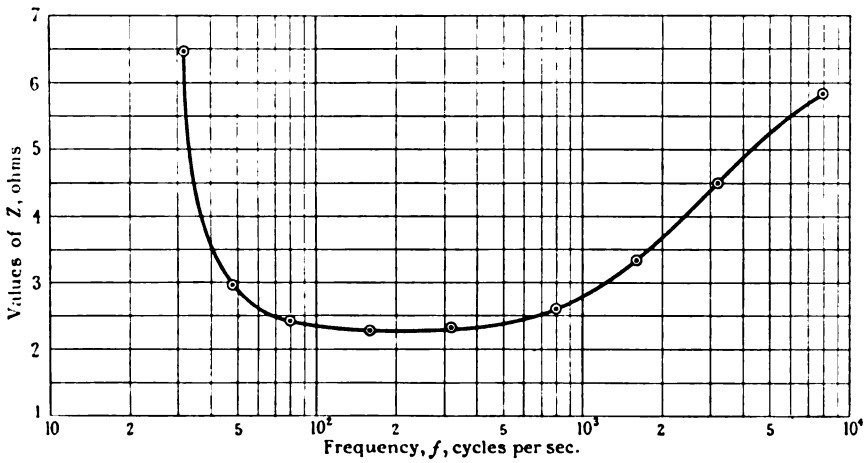

FIG. 17.—The re-computed electrical impedance of the loud-speaker.

Fixed Factors.

m = actual mass of moving parts = 21 grammes (mass of coil, former, etc. = 5 grammes).

A = force factor = $4 \cdot 28 \times 10^6$ C.G.S. units.

$D = A^2 \times 10^{-9} = 18 \cdot 3 \times 10^3$.

s = suspension stiffness = $0 \cdot 915 \times 10^6$ dynes per cm.

(9) RE-COMPUTATION OF INPUT IMPEDANCE.

These data are now utilized in Tables 6 and 7, firstly to compute the input impedance of the speaker and compare with the original measured values, and then to find the actual power input and output and response curve.

In Table 6 we first add to X'_{d_2} the reactance due to the 5 grammes of concentrated mass in the coil and former, and then subtract the stiffness reactance of the central washer: the result, in line (6), is X_d (in C.G.S. units). From X_d and R_d we find X_p and R_p by $X_p = D/X_d$ and $R_p = D/R_d$, which follow from equations (12) to (14). From X_p and R_p we find X_r and R_r by equations (25) to (28), and by adding X_c and R_c

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TABLE 6.

Re-Computation of Input Impedance.

(1)	$\omega = 10^3 \times$	0·2	0·3	0·5	1	2	5	10	20	50
(2)	$X'_{d_2} = 10^3 \times$	$6·6_5$	10·1	16·6	32·2	55·2	46	43	42	42
(3)	$5\omega = 10^3 \times$	1·0	1·5	2·5	5	10	25	50	100	250
(4)	$Sum = 10^3 \times$	7·65	11·6	19·1	37·2	65·2	71	93	142	292
(5)	$s/\omega = 10^3 \times$	4·58	3·05	1·8	0·9	0·4	0·2	0·1	—	—
(6)	$X_d = 10^3 \times$	3·07	8·55	17·3	36·3	64·8	70·8	93	142	292
(7)	$R_d = 10^3 \times$	0·36	0·78	2·53	11·3	45	112	121	124	125
(8)	$X_p =$	−5·95	−2·14	−1·06	$-0·50_5$	$-0·28_2$	$-0·25_8$	$-0·19_6$	−0·128	$-0·062_5$
(9)	$R_p =$	50·8	23·4	7·20	1·62	$0·40_6$	$0·16_3$	$0·15_1$	$0·14_7$	$0·14_6$
(10)	$k =$	−8·51	$-10·9_5$	−6·80	−3·21	−1·44	−1·58	−1·30	$-1·14_1$	−2·34
(11)	$1/k =$	−0·13	− 0·091	−0·15	−0·29	−0·69	−0·63	−0·77	$-0·87_7$	−0·42
(12)	$q =$	−8·64	$-11·8_6$	−6·95	−3·50	−2·13	−2·21	−2·07	$-2·01_8$	−2·76
(13)	$R_r =$	0·690	0·181	0·153	0·144	0·132	0·117	0·095	0·064	0·023
(14)	$X_r =$	−5·89	−1·97	−1·04	−0·465	−0·191	−0·073	−0·073	−0·073	−0·052
(15)	$X_c =$	0·04	0·07	0·11	0·22	0·46	1·07	2·00	3·25	3·90
(16)	$R_c =$	2·09	2·09	2·10	2·13	2·18	2·28	2·60	3·12	4·40
(17)	$X_2 =$	−5·85	−1·90	−0·93	$-0·24_5$	+0·269	+0·997	+1·93	+3·18	+3·85
(18)	$R_2 =$	2·78	2·27	2·25	2·27	2·31	2·40	2·70	3·18	4·42
(19)	$X_2^2 =$	34·22	3·61	0·86	0·06	0·07	0·99	3·72	10·1	14·82
(20)	$R_2^2 =$	7·73	5·15	5·06	5·15	5·33	5·76	7·29	10·1	19·53
(21)	$Z_2^2 =$	41·95	8·76	5·92	5·21	5·40	6·75	11·01	20·2	34·35
(22)	$Z_2 =$	6·48	2·96	$2·43_5$	$2·28_5$	$2·32_5$	2·60	3·32	4·50	5·86

TABLE 7.

Output Power.

(1)	$\omega = 10^3 \times$	0·2	0·3	0·5	1	2	5	10	20	50
(2)	$Z =$	6·48	2·96	$2·43_5$	$2·28_5$	$2·32_5$	2·60	3·32	4·50	5·86
(3)	$I =$	0·77	1·69	2·06	2·19	2·15	1·92	1·51	1·11	$0·85_3$
(4)	$F = 10^5 \times$	3·3	7·25	8·81	9·38	9·21	8·21	6·47	4·75	3·66
(5)	$R_d^2 = 10^6 \times$	0·13	0·61	6·40	127·7	2 025	12 770	14 640	15 380	15 620
(6)	$X_d^2 = 10^6 \times$	9·42	73·1	299·3	1 317·7	4 200	5 013	8 649	20 160	85 260
(7)	$Z_d^2 = 10^6 \times$	9·55	73·71	305·7	1 445·4	6 225	17 780	23 290	35 540	90 880
(8)	$Z_d = 10^3 \times$	3·09	8·59	17·5	38·0	78·9	133	152·5	188·5	301·5
(9)	$v =$	107	84·5	50·4	25·7	11·7	6·2	4·24	2·52	1·22
(10)	$R_u = 10^3 \times$	0·18	0·40	1·44	6·30	21	36	38	39	40
(11)	$W_2 =$	0·207	0·286	0·365	0·384	0·287	0·138	$0·068_5$	$0·024_8$	0·0059
(12)	$W_1 =$	3·85	8·45	10·3	$10·9_5$	$10·7_5$	9·6	$7·5_5$	$5·5_5$	$4·2_5$
(13)	$\eta\% =$	5·4	3·4	3·5	3·5	2·67	1·43	0·91	$0·40_5$	$0·13_8$
(14)	$P_{cr} \propto$	14·4	17	19·1	19·6	17	11·8	8·28	4·98	2·43
(15)	$P_{ci} \propto$	18·7	10	9·25	$8·9_5$	7·9	6·15	5·5	4·5	2·85
(16)	$dB_{cv} =$	−2·4	−1·2	−0·2	0	−1·2	−4·4	−7·5	−11·9	−18·2
(17)	$dB_{ci} =$	+6·4	+1	+0·3	0	−1·1	−3·3	−4·4	− 6·0	−10·9

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find X and R. Z is found in the usual manner, and is given in line (22) of Table 6, and also as a curve in Fig. 17. X and R are marked as points in diamonds on Figs. 4 and 5, to show the quite good agreement with the original data.

(10) Computation of Output.

Starting now with Z in Table 7, an e.m.f. $E = 5$ volts was assumed, to give the figures some concrete interest. It may be remarked that as the speaker is used with a 40/1 transformer this corresponds to 200 volts (R.M.S.), or, allowing for 20 per cent loss in the transformer, to 250 volts (R.M.S.) on the primary. The transformed impedance being of the order of 5 000 to 10 000 ohms, the " generated " e.m.f. in the valve anode circuit (with anode a.c. resistance of, say, 1 000 ohms) will be of the order of 250–300 volts, or 750–900 volts " swing." This would call for d.c. values of, say, 500 volts and 80 mA, or 40 watts dissipation.

From Z and E we find I, and hence F, in dynes. R_d and X_d give Z_d, in C.G.S. units, and from F and Z_d we get v, in cm per sec. Here again it is interesting to diverge from the main argument, and convert one or two values of v into U, the actual displacement. For example, at 80 cycles per sec., or $\omega = 500$, we have $U = v/\omega = 0{\cdot}088$ cm (R.M.S.), or $1{\cdot}25$ mm amplitude. At 48 cycles per sec., $\omega = 300$, $U = 4$ mm amplitude. This is about the limit of motion before the central washer causes distortion sufficient to be noticeable to a keen ear with the particular suspension in use, so that the input must be reduced for frequencies below 48 cycles per sec., unless one is prepared to tolerate the introduction of large percentages of harmonics—though, to judge by the tightness of suspension of many commercial loud-speakers, this is not found so objectionable by the average purchaser as it is by the author.

At $\omega = 2 \times 10^4$ (frequency 3 200),

$$U = \frac{2{\cdot}52}{{\cdot}2 \times 10^4} = 0{\cdot}000126 \text{ cm},$$

or an amplitude of $0{\cdot}0018$ mm, although the sound is painfully loud with this input power.

Returning to Table 7, line (11) gives v^2R_u, the output power, which has been converted to watts. v and R_u being in cm/sec. and C.G.S. units, the power appears in ergs, and is converted to watts on multiplying by 10^{-7}. This output power is denoted by W_2, and for interest W_1, the input power, is also computed so that the maximum power efficiency can be found. Excluding the case of 32 cycles per sec., where we are approaching resonance, we find $3{\cdot}5$ per cent at 160 cycles per sec. It will be noted that the maximum input power of 11 watts bears a reasonable relation to the estimated total d.c. dissipation of 40 watts, the ratio being $27{\cdot}5$ per cent.

Line (14) is proportional to the air pressure at a distance, and has been called P_{cv} as it corresponds to constant input voltage. If we divide each value of P_{cv} by the current which caused it, we get P_{ci}, which is the corresponding figure for constant-current operation. In lines (16) and (17) these pressures are converted into decibels up or down from the values for $\omega = 10^3$, and the resultant response curves are plotted in Fig. 18.

It is a source of great regret to the author that he has not been able to get a final check on the whole argument by having direct response curves taken.

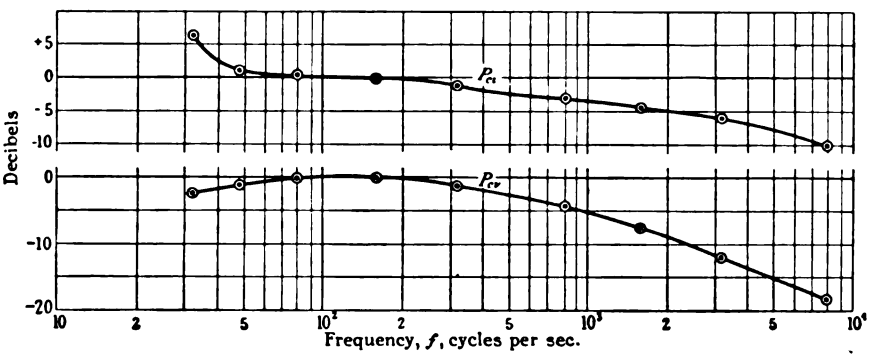

Fig. 18.—The computed response curves of output air-pressure, plotted against frequency. Referred to zero noise-level at 160 cycles per sec.

Owing to circumstances over which he has no control, it has been impossible for him to do so, but he believes that the particular speaker measured is still in existence, so that any enthusiastic volunteer with the necessary equipment may carry out the experiment if he cares to do so.

The author has used an almost identical instrument at home for a year. Speaking qualitatively, it is his belief that the response curve *is* substantially in agreement with Fig. 18. Although it is in no way germane to the present discussion, he cannot refrain from pointing out that this response curve is distinctly a good one.

(11) Conclusion.

The actual purpose of the present paper has been to call attention to the method of measurement, which is believed to be a really useful one. Unfortunately, the author has only been able to carry out the single series of measurements here dealt with. Although he conceived the method as far back as the spring of 1927, it was not until 1928 that the apparatus was available, and it was late in 1929 before he had at the same moment a suitable loud-speaker and time to do the work. Quite soon after this series of measurements had been carried

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out, a change of his firm's policy with regard to research in general effectually prevented any further attempts or any efforts to get a check by the direct measurement of output.

Considerable time was expended during 1930 in attempting a solution of the " line," before this was given up as being a task for a specialist in mathematics. The author has therefore been constrained to present this paper merely as an example in the use of a promising method.

Lastly, he wishes to apologize in advance for any clerical or numerical errors which may have escaped his vigilance. It is extraordinarily difficult to avoid such errors when both computation and checking have to be done single-handed, and on measurements which have also been made single-handed by the same worker.

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Discussion before the Wireless Section, 4th February, 1931.

Dr. G. F. Dutton: The author is to be congratulated upon so boldly tackling the loud-speaker from the electrical side. The cone has been treated as a simple line with one degree of freedom, whereas a cone or any thin sheet or diaphragm has two degrees of freedom. Can a diaphragm also be represented by a simple line? In the cone we have transmission along the radius from the apex to the circumference and also transmission round the cone at right angles to the radial transmission. It is due to these two degrees of freedom that we find the vibrations of the cone splitting up into nodal diameters and circles, and the efficiency at the high frequencies is due to the large number of natural modes of vibration, the fundamental of which is very low, say about 80 cycles per sec. The output response curve shown in Fig. 18 does not necessarily give the true acoustic pressure response in a confined space. The actual pressure at a particular frequency will depend upon the reflecting power of the walls of the room at that frequency. In order, therefore, to compare these curves with pressure response curves taken with a condenser microphone, the reverberation frequency characteristic of the room must be known. The curves given in Fig. 18 are quite unlike other loud-speaker

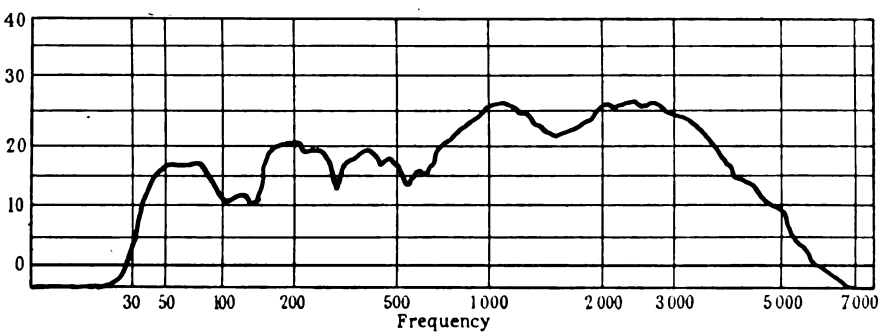

Fig. A.—Response curve for loud-speaker with 7-in. diam. buckram cone. Constant voltage across speaker.

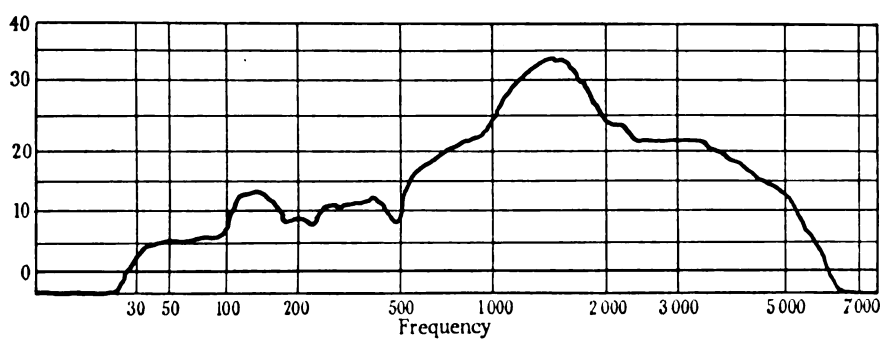

Fig. B.—Response curve for loud-speaker with 10-in. diam. paper cone. Constant voltage across speaker.

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response curves that I have seen, except perhaps when the speaker and transformer impedances have been mis-matched. Figs. A and B show the usual types of curves for a buckram cone (7 in. diam.) and a paper cone (10 in. diam.). The curves were taken at constant voltage, and they show the rise in efficiency between about 800 and 3 000 cycles per sec. I presume that the resonances of the nodal cycles and diameters are less damped with a paper than with a buckram cone. I should like to know upon what type of cone the author made his measurements, and whether there is any radial modes occur we have taken 300. This is due to the low-frequency resonances lasting only a few cycles. The results of one of our tests—in which not all the experimental points are shown—is illustrated in Fig. C, and it is clear that there are numerous resonances throughout the acoustic register. The author used a diaphragm of almost identical dimensions (apart from his surround) with that of Fig. C, but he says he neither obtained nor expected to obtain resonances. The curve of Fig. C obviously shows that the ramifications in the various mechanical and electrical coefficients are not in

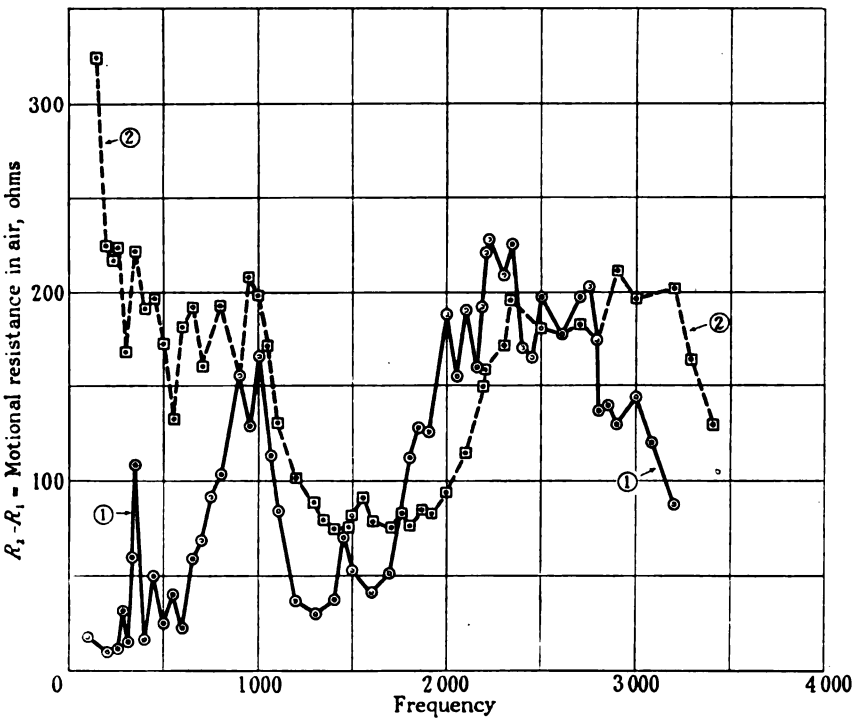

FIG. C.

Curve 1.—90° conical diaphragm 12·2 cm radius with reinforced edge, 1 000-turn coil.
Curve 2.—90° conical diaphragm 12·2 cm radius with rubber surround, coil 1·8 cm wide. The rubber surround acts as an auxiliary resonant diaphragm and enhances the lower register. The resonance of the surround *per se* occurs at 139·5 cycles per sec., whilst that of the diaphragm on the surround occurs at 18·7 cycles per sec. The ratio of these frequencies is 7:1, which is in good agreement with that calculated with the formula $f_1/f_0 = \pi(m/m_1)^{\frac{1}{2}}$.

possibility of taking a pressure response curve of that cone to compare with his computed values.

Dr. N. W. McLachlan: I hope to show by analysis of the results recorded in the paper, and with the aid of a selection from over 10 000 experimental observations on diaphragms of different size and thickness, made by Mr. G. A. V. Sowter and myself, that the author's observations and theory are alike unreliable. Mechanics teaches us that a vibrating system, like a conical diaphragm, must have resonances. No resonances are detectable in the experimental results, this being due to (*a*) the large frequency gaps between the readings, and (*b*) the curve-smoothing process adopted to overcome difficulties and mask physical realities. We have found it necessary, in order to ascertain the ultimate behaviour of a diaphragm, to take readings for at least 100 frequencies, and in cases of free-edge cones where keeping with the idea of a simple transmission line, as the author claims. A number of the resonances are quite prominent, e.g. 2 000 to 3 000 cycles per sec., and these should appear in the calculated output curve, whereas there is no sign of them in Fig. 18. In fact, owing to the concentration of energy at high frequencies into a beam, the resonances would be much more marked when plotted on an axial output pressure basis than as shown in Fig. C. A number of records of impulse tests on different kinds of loud-speakers have recently been published* which show serious resonances due to the diaphragm structure. In the case of an impulse, some of the resonances shown in Fig. C are heavily damped by the magnetic field, but the resonances to steady currents still remain. Turning to his numerical data,

* *Wireless World*, 1929, vol. 24, pp. 346 and 385; also *Philosophical Magazine*, 1931, vol. 11, p. 1.

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the author states that resonances were absent and that wave motion did not begin until 300 cycles per sec. was reached. Accordingly I have carefully analysed Table 1A with regard to two salient features:—(*a*) the accession to inertia, and (*b*) the efficiency as compared with that of a rigid disc. Two methods of calculation are used for the accession to inertia:—(1), in which the diaphragm loss is neglected, and (2), in which this loss is taken into account. The formulæ for these are as follows:—

$$(1)\quad M_1 = \left(M - \frac{s}{\omega^2}\right)\left[\frac{X_2 - X_3}{X_1 - X_2}\right]$$

where M_1 = accession to inertia,
M = mass of diaphragm, coil, etc.

$$(2)\quad M_1 = D\left[\frac{X_1 - X_2}{Z_m^2} - \frac{X_1 - X_3}{Z_{mv}^2}\right]$$

where $Z_m^2 = (R_2 - R_1)^2 + (X_2 - X_1)^2$
$Z_{mv}^2 = (R_3 - R_1)^2 + (X_3 - X_1)^2$

My results by both methods are in agreement with those of the author from 39·8 to 79·6 cycles per sec. At higher frequencies method (1) gives agreement, but method (2) does not. The results by method (1) show a constant value of M_1 (about 18 g) from $f = 0$ to $f = 159$ cycles per sec. From a physical viewpoint —owing to circulation waves from back to front—this cannot be the case with a 3-ft. baffle as used by the author. As the frequency falls from 159 cycles per sec., M_1 should also fall, and at 39·8 cycles per sec. it should be about 9 g, and not 18 g as shown in Fig. 13. In our own tests on a similar diaphragm having a reinforced edge, with a 6-ft. baffle *but no surround*, $M_1 = 12$ g at $f = 150$ cycles per sec., and $M_1 = 6$ g at 24·5 cycles per sec. This is in accordance with the theory of a rigid disc in a finite baffle, which is treated in my communicated remarks. Using method (2), the value of M_1 at 239 cycles per sec. is 14·3 g, whereas it should be about 18 g. The error lies in the values of R_1 and R_2. If an error of 1 per cent is made in either R_1 or R_2 the error in M_1 is 15 per cent. We have found that errors greater than this can occur due to a variety of experimental causes. One important factor is temperature, the resistance variation of copper being 0·4 per cent per deg. C. Since 40-minute intervals were required for the tests it follows that measurements of R_1, R_2 and R_3 were made at different times. It is well known that the heat dissipated in the magnet windings causes an appreciable rise in temperature, and it is therefore highly probable that a difference of several deg. C. will occur between the measurements of R_1, R_2 and R_3. A temperature difference of 5 deg. C. would entail an error of 2 per cent in R_2, and therefore of 30 per cent in M_1. In addition, since the radiation resistance is only 5 per cent of R_1 an error of 2 per cent in the latter means an error of 40 per cent in the former and also in the efficiency. In Fig. 6 it is assumed that the inductance of the coil is constant from 0 to 500 cycles, but this is not the case. In our tests the inductance of a certain coil varied from 238 μH at 500 cycles per sec. to 312 μH at 50 cycles per sec., i.e. a change of 30 per cent, which is not negligible. The efficiency of the author's loud-speaker is calculated from Table 1A using a formula recently given elsewhere.* The efficiency as calculated is 8·2 per cent at 39·8 cycles per sec., and falls to 4·2 per cent at 239 cycles per sec. The author (page 93) obtains the value 3·5 per cent at 160 cycles per sec., whereas I find it to be 5·2 per cent. The corresponding values on the rigid-disc theory using an infinite baffle are 1·8 per cent at 39·8 cycles per sec. and 1·7 per cent at 239 cycles per sec. How are we to account for these large discrepancies? In our observations on a number of diaphragms with reinforced edges and a 6-ft. baffle, but no surround, we did not obtain efficiencies in excess of those from the rigid-disc theory from 70 to 250 cycles per sec. When, however, the edge of the diaphragm was free, a series of sharp peaks giving efficiencies several times that of a rigid disc were obtained due to radial modes. Addition of a rubber surround smoothed out these peaks, but gave higher efficiencies than that from the rigid-disc theory owing to the surround acting as an auxiliary resonant diaphragm.† This must not be confused with the resonance of the diaphragm on the surround, since it occurred below 20 cycles per sec. We have shown‡ that, if

f_0 = frequency of diaphragm on surround (determined experimentally without centring washer),
f_1 = frequency of surround *per se*,
m = effective mass of moving system, including accession to inertia,
m_1 = effective mass of surround, including accession to inertia,

$$f_1/f_0 = \pi(m/m_1)^{\frac{1}{2}}$$

In the author's case this gives the surround frequency as about 8 times that of the diaphragm on the surround. Assuming $f_0 = 4$ cycles per sec., $f_1 = 32$ cycles per sec., which would have an effect upon the author's measurements. We have found both M_1 and the efficiency η to be increased by the resonant influence of the surround. Moreover, the increase in the values of these quantities found from the author's measurements, as compared with those for a rigid disc, is due either (*a*) to the influence of the surround or other resonance effects, e.g. the centring spider, or (*b*) to errors in the measurements themselves. I am of the opinion that it is caused by both. If the measurements were accurate and resonance were absent, the calculated values of M_1 and η would be in substantial agreement with the rigid-disc theory, using a 3-ft. baffle. Having shown that the moving-coil system is resonant throughout the frequency range, it follows that an analysis purporting to predict the acoustic output must take the resonances into account. The problem of prediction surely is to calculate the performance from given data. There should be no need for experimental verification as suggested by the author. His loud-speaker was demonstrated to me at Messrs. Graham Amplion's works some time ago; it had a powerful bass and a shrill upper register. The former

* *General Electric Review*, 1930, vol. 33, p. 507.
† See *Philosophical Magazine*, 1931, vol. 11, p. 2.
‡ *Philosophical Magazine*, *loc. cit.*

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appears in his measurements and is probably due to the surround acting as an auxiliary resonant diaphragm. The latter is not evident in Fig. 18.

(*Communicated*): At very low frequencies, when the baffle is small compared with the wavelength of sound, the accession to inertia of a vibrating disc can be derived from the solution to the more general problem of an ellipsoid vibrating along one of its axes. In view of the importance of the subject in modern acoustics, it may not be out of place to present the matter in detail. In 1833 Green,* whilst considering the aerial forces on a pendulum bob, showed that the accession to inertia of an ellipsoid vibrating along the x-axis with no baffle is given by

$$M_1 = \rho v I/(2 - I)$$

where $I = abc\int df/a_1b_1c_1$

$(x^2/a^2) + (y^2/b^2) + (z^2/c^2) = 1$

$a_1^2 = a^2 + f$
$b_1^2 = b^2 + f$
$c_1^2 = c^2 + f$
} confocal conicoids

(a, b, c, being functions of x, y, z, determined from these equations)

ρ = density of air under given conditions of temperature and pressure,

and v = volume of ellipsoid.

This result is applicable provided that the axial amplitude is small and the sound waves long compared with the

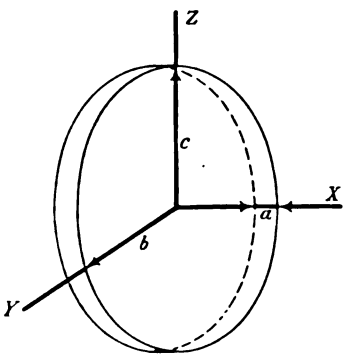

FIG. D.

Spheroid vibrates along axis of X. $b = c >> a$

dimensions of the ellipsoid. When $b = c$, and a is small in comparison, the spheroid is substantially a disc of thickness a, as shown in Fig. D.

Then

$$I = ab^2\left\{2p^{-\frac{1}{2}}\left[\arctan\left(ap^{\frac{1}{2}}\right) - \pi/2\right] + 2a^{-1}p^{-1}\right\}$$

where $p = b^2 - a^2$
$= b^2$, since $a^2 \to 0$.

Since $a << b$,

$$I = 2 - \pi a/b$$

and $M_1 = \rho(4/3)\pi ab^2(2 - \pi a/b)/(\pi a/b)$
$= (8/3)\rho b^3$, where b is the radius of the disc.

* *Transactions of the Royal Society of Edinburgh*, 16th Dec., 1833.

Rayleigh* has shown that the value of M_1 for both sides of a disc in an infinite baffle is $(16/3)\rho b^3$. Thus, provided the specified conditions are satisfied, the effect of an infinite baffle is to double the accession to inertia. Moreover, with a baffle 3 ft. square at very low frequencies (including zero) the accession to inertia of a disc 12 cm radius is about half its value at 200 cycles per sec. If $a = b = c$, $M_1 = (2/3)\pi b^3$, this being the result for a sphere. Thus the ratio of the values for the disc and the sphere, in the absence of a baffle in both cases, is $8/(2\pi) = 1{\cdot}28$. This shows that M_1 is influenced by the shape of the diaphragm. The power radiated by a diaphragm vibrating in a finite baffle has been investigated by M. J. O. Strutt.† He uses a model consisting of two single sources in anti-phase at the extremities of a spherical diameter. The half-circumference of the sphere is equal to the side of the baffle, which is assumed to be circular. If c = radius of sphere, and a = radius of baffle, $2a = \pi c$, or $c = 2a/\pi$, this being the radius of the spherical model. The object of such a model is, of course, to make the problem yield readily to mathematical analysis. Strutt gives a curve from which the output can be calculated at any frequency for any size of baffle. In using the curve it must be borne in mind that the mathematical source is very small compared with the baffle, which is seldom the case in practice.

Mr. C. G. Garton: I agree with Dr. Dutton that it is impossible to represent a cone by a simple transmission line. This is clearly shown by experiments with powder distributed over the cone. The type of vibration observed does not at all resemble a simple wave system travelling down the cone and compressing it into a series of zonal corrugations. It is rather a system of small isolated islands of vibration produced by the combination of two or more sets of oscillations both in the direction of the axis and circumferentially. I do not agree with Dr. McLachlan's statement that the increase of amplitude in the upper register over that which one would theoretically expect is due entirely to flexure of the coil. The extent of the upper register in loud-speakers of this type depends, to a large extent, upon the material of the cone and the flexural losses in the latter. If a very soft paper (e.g. blotting-paper) is used the cut-off occurs at about 2 000 to 3 000 cycles per sec. If harder paper is employed the corresponding frequency is 5 000 to 6 000, whereas by using a material with very much lower losses (e.g. duralumin or aluminium) one can extend the register to well over 10 000 cycles per sec. This is a proof that it is the material, in addition to the actual construction, which affects the higher register. While congratulating the author upon the amount of work that has been put into this investigation, I am doubtful whether much progress will be made in the design of this type of loud-speaker as the result of such measurements. It seems to me that the problem of the cone vibration is far too complex for analysis by means of measurements. Although an accurate analysis can, it is true, be applied at very low frequencies, these are unfortunately not very important, so that in the whole of the useful range one is dealing with quantities which are analytically indeterminate.

* "Theory of Sound," vol. 2, 2nd ed., p. 165, expression 14.
† *Philosophical Magazine*, 1929, vol. 7, p. 537.

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I therefore think that, unfortunate though it is, we have to work from a purely empirical standpoint, and the difficulties are great. Some empirical work on which I have been engaged for several years leads me to believe that success will be achieved by the use of a far more rigid material than paper, and that the metallic-diaphragm type will replace the existing paper-diaphragm loud-speaker. By the use of thin duralumin (3 or 4 mils thick) it is possible not only to extend the register to 10 000 cycles and beyond, but also to obtain much more faithful reproduction, because of the better response to transients. Besides obtaining a steady-state response curve we have to obtain the curve of response of an impulsed cone, i.e. the cone must respond to transient forces. A paper cone does not do this satisfactorily, chiefly because it gives no output at the equivalent frequency of the transient. A transient can be considered to have an equivalent frequency well in excess of 5 000. The cone has no output in this region, and therefore if the transient is analysed into a Fourier series all the components in the higher part of the register will be found to be missing. Owing to the higher speed of sound in the metal cone, this responds better over the higher register. While a paper cone and a metal cone may have approximately similar static characteristics, it is found that the metal cone actually gives better reproduction of speech and other noises involving transients. When the static curve is continued to 10 000 cycles per sec. and beyond, a still further improvement is obtained. These results have been achieved empirically, and I think that future progress will be along these lines.

Mr. D. A. Oliver: I should like to congratulate the author upon his bold attempt at measuring the total watts output of a loud-speaker. Here is the first detailed attack by this new method, and as the differences in effective resistance which are involved are so small it would perhaps be worth while to look into the possible accuracy of the method. I should like to ask the author whether he considers that the degree of vacuum achieved in his experiments was sufficiently high to remove entirely the acoustical load on the diaphragm. I believe that with the pressure used by the author a much larger chamber is necessary in order that the mean free path conditions for the removal of the load may be satisfied. The wave transmission in the cone is an interesting problem, and my own opinion is that for " concertina " action, i.e. where there is purely circular symmetry about the axis, the cone can be represented by a transmission line. I do not think, however, that the constants of the transmission line can be obtained by the method described in the paper, on account of several important factors. For instance, this method relies upon the assumption that the driving-point mechanical impedance of the cone itself remains the same in the vacuum as it is under the conditions of normal operation. It is unlikely that, when a distributed acoustical load is removed from a paper cone, the inherent mechanical impedance of the cone will, at a given frequency, remain unaltered. I hope to try for certain simple modes of vibration a method which is analogous to the series of operations employed in telephone engineering, namely short- and open-circuiting the periphery of the cone; first by having a perfectly free edge, and secondly by clamping the edge rigidly and making observations at the driving point, i.e. at the coil, which would be made as rigid as possible. I think that if this were done in air the main transmission-line constants for certain symmetrical modes could be obtained. As the author points out, it is unlikely that the cone system could be represented by a uniform line, and, this being so, I think that the further information which would be required comprises the vector ratio of the velocities at the inside and outside edges of the cone when the periphery is unclamped. These measurements having been obtained, the constants of the equivalent non-uniform line can be found, with the limitation that these do not hold if any radial modes of vibration exist. The falling-off in the output response shown in the curves of Fig. 18 is unusual, as curves of measured values almost invariably rise, especially in the region between 2 000 and 4 000 cycles per second.

Mr. G. A. V. Sowter: In connection with the bridge measurements, the author stresses the facts that (*a*) the source of supply was a beat oscillator designed with much care, (*b*) the amplifier was a 3-valve one of considerable power, and (*c*) the calibration of frequency was carried out by means of a Campbell frequency bridge and Western cathode-ray oscillograph in conjunction with an auxiliary oscillator. I suggest that had the author taken as much care over the conditions of test as with the details of the measuring equipment, he would not have to suggest " that he might be masking the symptoms of real effects." For example, it is stated that the moving coil was fixed in the pot with chocks and 2 kg of sand was poured into the cone. From much experience of this type of measurement with Dr. McLachlan I should point out that (*a*) The coil was not fixed, otherwise there would not be the change in reactance between 35 and 200 cycles per sec. indicated in Table 1. A pronounced dip is measured at 103 cycles per sec., indicating a resonance which must be due to motion of the coil. The use of chocks may also deform the coil sufficiently to introduce errors in R_1 which, although small, are of importance when the difference $R_2 - R_1$ is taken. Owing to the flexural modes of the coil at the upper frequencies, the chocks would be inadequate to prevent small amplitudes of the coil, thereby giving erroneous values of R_1. (*b*) Even if the coil were fixed, it is a matter of extreme difficulty to fix it in the same position as for the " normal " or free test. Experience has shown that a small change in position of the coil has an appreciable effect upon the measured values of R_1, R_2 and R_3, and since this is a method of differences very large errors are introduced in this manner. The effect of temperature is also of great importance. Referring to Table 1, it will be seen that the differences between the normal values of R_2 and the values of R_2 *in vacuo* are of the order of 2 to 3 per cent only. The magnet pot is energized from a supply of probably 20 to 30 watts, which causes a considerable rise in temperature of the magnet and moving coils. Moreover, since some observations took as long as 40 minutes to make, the two series of results could not have been obtained on the same day, and the observations might have been made at different temperatures.

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To effect a change of resistance of 2 to 3 per cent a temperature change of 6 to 7 deg. C. is all that is necessary, and if the temperature of the moving coil during the test was not known accurately to within fine limits the values of the measurements *in vacuo* are useless. Furthermore, *in vacuo* the heat from the magnet winding is not so readily dissipated as in air, and the temperature will tend to rise more rapidly, thereby vitiating the results. Equable temperature is surely far more important than some of the " essentials " of the bridge equipment. To be strictly accurate, in the measurements of accession to inertia the temperature and barometric pressure of the atmosphere should have been observed, since the density of air is involved. The author uses a square baffle of 3 ft. side, and takes measurements of accession to inertia in the neighbourhood of 50 cycles per sec. Does he know that below about 200 cycles per sec. the effect of the baffle falls away with decrease in frequency? It is hopeless to expect agreement between theory and practice unless the size of the baffle is taken into account. It is difficult to understand the reference to the microphone effect, because although it is stated (page 80) that the experiments were carried out in a room the effect of the wind seems to enter into the problem. The author evidently employed much more amplification than was useful, and this may have introduced difficulties. In addition, the type of vibration galvanometer used was unsuitable owing to its low damping and limited range. There are a number of vibration galvanometers on the market having frequency ranges of 20 to 200 or 30 to 300 cycles per sec. whose cost is a fraction of that of the rest of the equipment used by the author. Measurements involving the use of a telephone ear-piece at 150 cycles per sec. are in my opinion not of great accuracy, even with an amplifier. The sensitivity of the ear in this region is so small that large out-of-balance currents are required, and since harmonics are present (due to the iron in the circuit) even with a good oscillator, there is a tendency to take the balance due to harmonics rather than to the fundamental. This follows from the fact that the sensitivity of the ear increases up to 2 000 cycles per sec., and the output of the telephones increases with frequency up to their resonance point. The author states that the resonance of the centring washer of the loud-speaker, calculated from the measured stiffness, occurs at 32 cycles per sec., and it would be of interest to find this resonance experimentally. It can be done by a bridge method or by finding the frequency at which minimum current is obtained when the loud-speaker is in a valve circuit, a constant voltage of variable frequency being applied to the grid of the valve. I suggest that in taking the frequency intervals cited in the paper the author missed the important features of the moving-coil loud-speaker. His results below 300 cycles per sec. are not of great value, for the reasons mentioned, and above this frequency the intervals between measurements are much too great to determine the actual performance of his speaker. Moreover, he discards all measurements above 800 cycles per sec. as only suitable for making speculative inferences. The only method of determining the practical performance of a speaker by measurement is to take some hundreds of readings at close frequency intervals and concentrate on those frequency regions wherein resonances become apparent. Finally, the Bibliography is conspicuous by the absence of reference to the extensive work of Dr. N. W. McLachlan. He outlined the rigid-disc theory in his book " Wireless Loud-Speakers " in 1927, and in two papers* published the same year. These, together with a more recent paper,† completely anticipate Reference (2) in the Bibliography.

Mr. A. G. Warren: The author on page 82 attempts to justify upon theoretical grounds the smoothing of his curves of observed values. Throughout the paper there are references to the " results of analysis," " the fundamental equations," etc. These, however, are not given. Beyond a little elementary mechanics and electricity (headed " The Basic Analysis "), which bears but a very limited application to the problem in hand, no theoretical consideration is attempted. If the author has the analysis or the fundamental equations he could greatly assist by quoting them, for they are not generally known. The very limited analysis which has been published will not justify

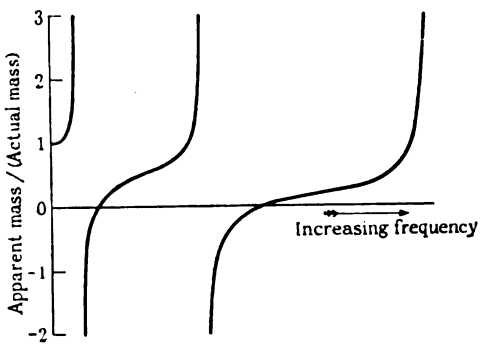

FIG. E.

any smoothing. Certain features of the phenomena are well known. In a diaphragm we have a structure with an indefinite number of modes of vibration. (The author observed his own diaphragm breaking up at 300 cycles per sec., and this process must be recurrent throughout the musical scale.) The free natural symmetrical vibrations of a flat diaphragm having no losses were worked out by Rayleigh. For these, there is motion of the centre with no force; the mass is zero, and it is still very small when losses are taken into account. The fixed centre vibrations are also of importance. For these, the mass is infinite, and is still very great when losses occur. I published the frequencies of these vibrations, and also the apparent mass for the forced vibrations, in May 1930.‡ Fig. E shows roughly what occurs. As the frequency rises the effective mass exceeds the actual mass at an increasing rate. After fixed-centre resonance is passed the effective mass becomes negative and then proceeds through zero (free resonance) to a high positive value, and so on. Steep peaks and ravines are to be expected in experimental curves. One naturally looks for them, because the resonances of which they are evidence are responsible for the high-frequency output of the loud-speaker.

* *Wireless World*, 1927, vol. 20, p. 372, and vol. 21, p. 357.
† *Philosophical Magazine*, 1929, vol. 7, p. 1011.
‡ *Ibid.*, 1930, vol. 9, p. 881.

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Fortunately they are present, though some of them are over in a range of but a few cycles. In Fig. 7 the author depicts what he considers to be an approximation to the line equivalent of a diaphragm. In a few simple cases it is possible to translate a mechanical problem involving alternating quantities into an electrical problem. It is not thereby intrinsically rendered any easier; all that is done is to express it in a language with which electrical engineers are generally more conversant. Unfortunately, however, electrical language is much more limited than mechanical, as many mechanical quantities have no electrical analogues. A velocity may be represented by a current, a mass by an inductance, a compliance by a condenser, but how is one to represent a bending moment? It may be represented by a formula, but the whole purpose of the diagram is vitiated if formulæ become the circuit components. One difference between a diaphragm and any known form of line is apparent at once. A continuous line has no frequency recurrence, a loaded line has a frequency recurrence but is not continuous, a diaphragm is continuous and had a frequency recurrence. Dr. Dutton has mentioned that a diaphragm has two degrees of freedom, i.e. it resonates with both circular and diametral nodes. This is an extra complication, but sufficient difficulties are encountered if only the symmetrical modes are considered. I have worked out the parameters of the line for the simple case of a flat diaphragm without losses. They are represented by complicated functions of Bessel functions which can convey no physical picture in terms of any known electrical elements. To attempt any deduction in terms of practical lines is only to invite error. The result is apparent. For instance, the author on page 91, referring to R_{d_2} and X_{d_2}, states that these quantities, being the components of the input impedance of the line, tend towards constancy at high frequencies. They do, but the mechanical quantities they are supposed to represent do not. This is one of the reasons why the idea of an equivalent electrical line must be rejected. The author states that he conceived the method of experiment as far back as the spring of 1927. By that time a considerable amount of work had been done; in particular, N. W. McLachlan had published the theory of the rigid-disc speaker, and the general lines along which research should be pursued had been indicated.

Mr. N. Fleming: I am unconvinced that the computed curves given in Fig. 18 for the acoustic response of the author's loud-speaker are at all a true picture of the actual response of the instrument. My lack of conviction is partly due to the author's method of treating the experimental results, and partly to some doubt which I feel as to the validity of the fundamental assumption upon which the calculation of acoustic power output is based. In the first place, the author makes measurements at a few widely spaced frequencies and then smooths the experimental values. Dr. McLachlan has already pointed out that the motional resistance and reactance of such an instrument show very large and sudden variations with frequency. For instance, in some measurements that I made about 4 years ago on a coil-driven cone-type instrument the motional resistance changed some sixfold between 640 and 700 cycles per sec. If measurements are made at widely spaced frequencies, and such irregularities as then appear are ignored, the true characteristics of the instrument are, to a large extent, lost. Using these smoothed results for frequencies up to 800 cycles per sec. only, the author calculates certain mechanical constants of the system. Above 320 cycles per sec. he further smooths these values and then extrapolates to 8 000 cycles per sec. Is is from these figures that he finally derives the acoustic power output. Since, on the basis of the assumptions he makes, the acoustic power output may readily be calculated directly from the experimental results, there appears to be no need for all this smoothing, except in so far as it leads to possible extrapolation. The author himself describes the latter as speculative, but I feel that, since it completely ignores the large irregularities in the measurements at high frequencies, it may be quite misleading. In order to obtain some idea of the effect of the smoothing, I have re-calculated the acoustic power output at various frequencies directly from the experimental figures, and in so doing have discovered a few slight errors. The values for R_2 and R_3 at 6 000 cycles per sec. given in Table 1 do not agree with the points plotted in Fig. 4. They are 4·0 ohms in one case, and 4·2 in the other. In line (1) of Table 7 the values of ω should presumably be the same as those in line (1) of Table 6. In Fig. 18 the points plotted at 700 cycles per sec. should be at 800 cycles per sec.,* and in the footline the figure should be "160" instead of "1 600."* Re-calculating from the experimental figures between 200 and 1 000 cycles per sec., I obtain deviations up to nearly 5 decibels from the curve for P_{cv} in Fig. 18. Above 1 000 cycles per sec. the author gives no experimental values for the "held" resistance and reactance of the coil, although he states on page 91 that the measurements extended up to nearly 5 000 cycles per sec. I think that it would be useful to include these measurements in Table 1. For the held impedance I employed, therefore, values taken from a curve through the figures given in Table 5. At 3 000 cycles, where the author's measurements lie well away from his smoothed curves, P_{cv} comes out about 10 decibels higher than the value given in the paper. Such a result would be in accordance with a peak in this region, and it has already been pointed out that this is a normal characteristic of a loud-speaker of this type. At 1 500 and 2 000 cycles per sec., however, the experimental results lead to negative values of the acoustic power output—an impossible result. It may be that the figures I have taken for the held impedance are incorrect, that the experimental results are incorrect, or, as Mr. Oliver has already suggested, that the fundamental assumption upon which the calculation of the acoustic power output is based is invalid. This assumption is that, for a constant driving-point velocity, the power absorbed internally in the system is the same whether the instrument is operating *in vacuo* or in air, so that the extra power dissipated in the second case represents the work done on the air. It seems to me quite possible that the reaction of the air upon the diaphragm may so change the mode of vibration of the latter that the power dissipated internally will not be the same in the

* Since corrected for the *Journal*.

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two cases. Consequently the acoustic power output cannot be inferred from their difference. As a check on my method of calculation, which differed from that adopted by the author, I took in one case ($\omega = 0{\cdot}5 \times 10^3$) the smoothed values given in Table 1A. I was rather surprised to find that the value of P_{cv} was not the same as that calculated by the author, but on going through his calculations I came to the conclusion that the error was his, and that the value of P_{cv} at 80 cycles per sec. should be $-0{\cdot}2$ decibel. It is perhaps of interest to notice that when this point is corrected on the curve of Fig. 18, and the 800-cycles point is brought into its correct place, the last vestige of irregularity is removed from the response of the author's loud-speaker.

Mr. A. M. Hallawell: The paper is interesting, not so much on account of the results achieved in the experiments described, but because it indicates a new line of attack upon a difficult problem. I think that the author would have been better advised to employ for his first experiment a speaker having a horn and an acoustic transformer, i.e. a sound-box. In such instruments the loading on the diaphragm can be relatively much higher than on the open type of cone speaker, and I think that had the horn type been used the influence upon the final results of experimental inaccuracies would have been less marked. I do not agree with Mr. Garton that this work is unlikely, in any event, to lead to valuable results; an important matter which still requires investigation is the question of standardizing the methods of measurement and specification of performance of loud-speakers. All the methods used at present are rather arbitrary. It is well known that the response curves commonly used at present depend upon where the measuring microphone is placed, how the pressures are averaged, etc. If the author's method were perfected it could be used to indicate the total acoustic output obtained for the diaphragm for a given input, a rather more fundamental characteristic than the pressure measurements which are generally employed at present. With regard to the discrepancy which has been pointed out between the falling characteristic obtained by the author and the rising characteristics usually obtained by pressure measurements, the pressure response curves usually shown are taken on or near to the axis of the speaker; if a response curve is taken at the side of a speaker it is quite possible that a falling characteristic may actually be found. Unless the responses at various angles all round the speaker are measured and averaged, it is difficult to say whether the total output should be represented by a falling or a rising curve. I disagree with the author's statement that the room in which the diaphragm works has very little influence upon the diaphragm loading. If a speaker which is emitting a constant note is moved only 1 ft. from its original position in a room, one can frequently hear that the mean intensity being emitted has changed, quite apart from any alteration due to a change in the relative position of the observer and the standing-wave pattern. This indicates definitely that the loading on the speaker must have altered, which would not be the case if the speaker were moved when in the open air.

Mr. M. G. Scroggie: The author points out that at the higher frequencies—800 to 1 000 and upwards—the results he obtains with the cone and coil held approach very closely to those with the cone and coil moving *in vacuo*. I am not certain, however, that the theoretical reasons he adduces for this effect are entirely responsible, and am particularly doubtful about his method of holding the coil. At 3 200 cycles per sec. the amplitude of the centre of the diaphragm is only of the order of 0·001 mm with full input to the coil, and even at 1 000 cycles per sec. it is only about 0·018 mm. My own experience is that it is extremely difficult to clamp a coil effectively, and I am sure that some motion of the coil, if not of the cone, is bound to take place if the only restraining force is 2 kg of sand poured into the upturned loud-speaker. Only a small proportion of the weight of sand bears on the coil itself, and it is difficult to believe that the coil does not clear a space for itself of the order of 0·001 mm high. The author points out that the response curves of Fig. 18 are distinctly good, but I think that he has helped matters considerably by the choice of the vertical scale, which is rather a cramped one. After making allowance for this, however, the absence of resonances, which may be due to the comparatively small number of observed values, has to be borne in mind. I cannot agree with him that the effect of standing waves is entirely confined to the usual type of measurements of the acoustic output of a speaker. It seems to me that the standing waves in the room must affect the impedance, and consequently the curves shown in Fig. 18. These, incidentally, approximately correspond to operation with a pentode valve (upper curve) and a triode valve (lower curve), i.e. to constant-current and constant-voltage conditions respectively. Neglecting the single reading at the lowest frequency, the falling-off at the upper frequencies in the constant-current curve is much less than that in the constant-voltage curve, which corresponds to the general qualitative results obtained with a pentode as compared with a triode valve. In spite of the remarks of previous speakers, tending towards adverse criticism of the validity of the author's measurements, the curve of Fig. 17, which shows the re-computed electrical impedance of the loud-speaker, does correspond very exactly in shape to curves plotted from direct measurements of large numbers of moving-coil speakers.

Dr. L. E. C. Hughes: The author has introduced a mechanical line to correspond to the action of a coil-driven cone. If the mechanical line is terminated by the correct impedance (the surround) to ensure uniform motion along the cone radius for all frequencies, the resonances due to the diaphragm breaking up into segments are eliminated. Dr. McLachlan* has, however, shown that these resonances are essential for a reasonable efficiency at high frequencies. The prominent resonance at about 3 000 cycles per sec. which occurs in most cone speakers may be substantially reduced by adding a narrow ring of paper half-way along the radius, thus breaking the uniformity of the standing wave. The author's vacuum method may prove useful in the case of the mechanical analysis of more efficient loud-speakers, but the overall response of a loud-speaker consists of three separate responses—electrical, mechanical and acoustic. The electrical

* *Philosophical Magazine*, 1931, vol. 11, p. 1.

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response may be regarded as the power reflection-loss due to change in impedance, the mechanical response involves the mechanical line with its resonances and the acoustic response involves the amount of sound

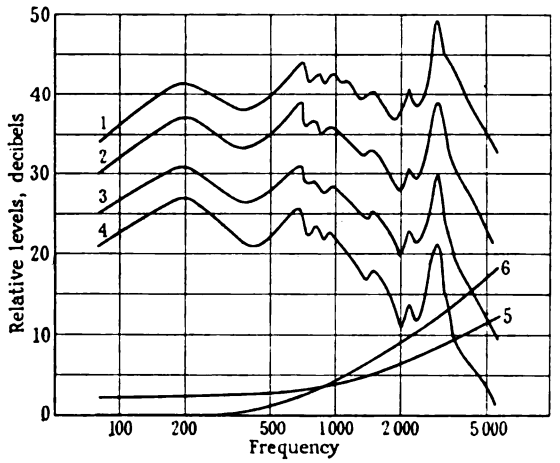

FIG. F.—Coil-drive loud-speaker response curves.

power radiated, together with its distribution in space. The latter has recently been studied extensively by Stenzel* and Wolff-Malter,† both theoretically and practically, but although the directivity of loud-speakers has been appreciated for a long time, it remained for McLachlan‡ to calculate the distribution of the sound field radiated from a piston operating in a plane, and to give polar diagrams for different frequencies. The publication of these curves explained results which I obtained as a student with Prof. Mallett. The response curves of three different loud-speakers were obtained by placing a Rayleigh disc on the axis of the loud-speaker in a large padded chamber. The result for the case of a constant-current input to a coil-drive loud-speaker is shown in Fig. F by Curve 1. In order to obtain the total sound-power output it is essential to find the polar distribution of the radiated intensity, and to integrate this by Rousseau's construction.* Three such curves, for 192, 2 000 and 3 300 cycles per sec., are indicated in Fig. G by the Curves 7, 8 and 9, plotted on a radial decibel scale. For comparison, McLachlan's curves (Curves 11 and 12) for 1 024 and 4 096 cycles per sec. have been added. At low frequencies there is theoretically uniform distribution (Curve 10), while at high frequencies the negative loops appear to be negligible acoustically, being about 17 decibels below the axial intensity at 4 096 cycles per sec. The agreement between the experimental polar diagrams and the theoretical diagrams for an equivalent piston is clearly shown by drawing the curve for the concentration factor. As this is defined as the ratio between the axial and the mean hemispherical intensity, the concentration factor may be regarded as the axial "gain" of the distribution; it is shown theoretically (as calculated from the curves given by McLachlan) in Fig. F by Curve 6, and experimentally by Curve 5. When Curve 5 is deducted from Curve 1, Curve 2 is obtained. This gives the total sound power radiated for (constant-current)^2 input.

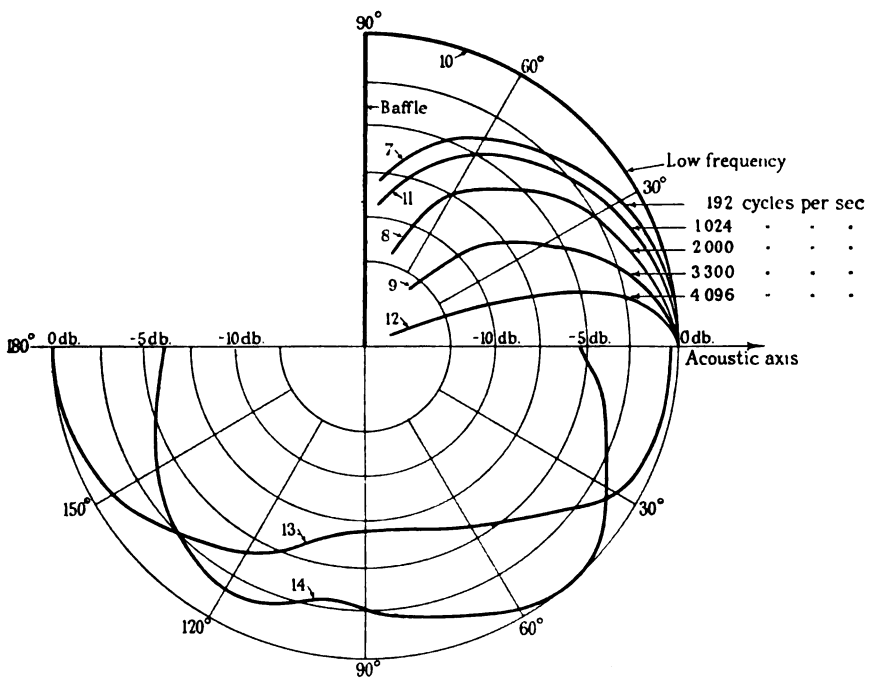

FIG. G.—Loud-speaker polar curves.

Curves 7 to 12—Coil-drive loud-speaker.
Curves 13 and 14—" Kone " loud-speaker.

* *Elektrische Nachrichten-Technik*, 1927, vol. 4, p. 239, and 1929, vol. 6, p. 165.
† *Journal of the Acoustical Society of America*, Oct. 1930.
‡ *Wireless World*, 1927, vol. 21, p. 357.

* D. A. OLIVER: *Experimental Wireless*, 1930, vol. 7, p. 653.

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Curve 3 is obtained by deducting $10 \log_{10}$(resistance) from Curve 2, and hence shows the relative conversion efficiencies of the loud-speaker. The question arises, What curve is to be taken as *the* response curve of a loud-speaker? The author commences by quoting Bostwick,* but in equation (48) he departs from Bostwick's criterion for the response of a loud-speaker. There is no doubt that the correct response is the gain of the total sound power radiated over the electrical power which would be delivered to the loud-speaker if the impedance of the latter were a resistance which matched the output impedance of the source. This conception of the response automatically takes account of the impedance variations and the distribution of the radiated sound, and comes into line with the responses of other acoustic apparatus. It is particularly useful with irregular distributions, such as with the "Kone" loud-speaker, of which two polar curves are shown in Fig. G. Curve 13 refers to 180 and Curve 14 to 1 810 cycles per sec. The desired response curve (Curve 4) of the coil-drive loud-speaker is therefore obtained by assuming a constant electromotive force in series with an output impedance suitable for the operation of the loud-speaker. The correct response curve is thus very different from that taken along the acoustic axis. The efficiencies calculated by the author and by E. D. Cook† (who also used a vacuum method) are abnormally high for this type of loud-speaker. The figure I obtained as a maximum was about 0·4 per cent, the order of which has been confirmed by E. J. Barnes.‡ The problem of testing loud-speakers has been recently reviewed by Kellogg,§ and the problem of the microphone, which involves just as many difficulties, by W. C. Jones.‖

Mr. L. C. Pocock (*communicated*): The author has carried out an interesting experiment which many investigators must doubtless have wished to make. It is, however, the first experiment of the kind and, while commending the author's skill and patience, it would be unwise to accept without considerable reserve all the conclusions that are drawn. It must be remembered that the quantities of real interest are not measured directly, but are derived by means of hypothesis and theory from impedance measurements. In such a case corresponding experiments on a number of loud-speakers are necessary to confirm the theoretical basis of the calculation before it can be accepted as correct. Speaking broadly, the paper opens the way to a hitherto unexplored line of attack, and then passes to a re-calculation of the impedance data from the constants that have been derived. This appears to be like travelling in a circle, and is the slenderest of justification for finally extrapolating and deriving a performance characteristic. I think some caution should be exercised in accepting it as proved that wave propagation begins at as low a frequency as 500 in conical diaphragms of relatively small size. The sand figures obtained on double-cone diaphragms¶ indicate, and theory supports, that sector vibration sets in at lower frequencies than nodal circle vibration, and one experimenter* at least, observing mirrors fixed on the diaphragm of a Rice-Kellogg loud-speaker, asserts that it vibrates as a whole. There appears to be scope for further work upon the mode of vibration of small paper cones and, in particular, upon the reaction of various modes of vibration upon the driving-point impedance. The author bases a good deal upon the calculated air impedance for a circular disc in a baffle. I would point out that the resistance curve in Fig. 1 should oscillate appreciably with decreasing amplitude about its asymptotic value. H. Stenzel has published† a thorough theoretical study of the directional and other properties of a radiator which is an approximation to a cone, some of his conclusions being substantiated by direct measurements on cones. The principal conclusion which he derives is that the conical shape of the radiator introduces a number of deep absorption troughs and successively diminishing resonant peaks, beginning at about 2 000 cycles per sec. in the Rice-Kellogg loud-speaker. Further, it has been found in a number of measurements made on loud-speakers in the International Telephone and Telegraph Laboratories that a fairly wide absorption trough in the middle frequency range is caused in practice by the finite size of baffle used, the location of the trough shifting with changes in baffle dimensions. It is therefore rather unlikely that the upper part of the calculated frequency characteristic in Fig. 18 could be verified by direct measurement.

Mr. P. K. Turner (*in reply*): Before replying in detail to the discussion, I wish to call attention to a misapprehension which seems to exist in the minds of some of those who have spoken. The paper is essentially a description of a method of measurement: the results of a test carried out by the method were given simply as an example; and errors (if any) in the measurements or in the inferences drawn from them must not be taken as invalidating the method.‡

The observations in the discussion seem to me to fall under three headings: (*a*) criticisms of the method, (*b*) criticisms of my application of it, and (*c*) observations on the theory of loud-speaker action which, however interesting, are irrelevant. I propose to deal first with points raised as to the soundness or otherwise of the method.

Dr. Dutton states that my response curve does not necessarily give the true acoustic pressure response in a confined space. I certainly hope not, for the response in a confined space is probably as much dependent on the properties of the space as on those of the speaker; and I want as far as possible to get a curve which is a function of the speaker only—this very point is one of my chief objections to the usual type of response curve.

Mr. Oliver asks whether the degree of vacuum was sufficient. The remaining air was a few thousandths of the normal amount: hence all but a few thousandths of its effects were removed, which gave ample accuracy for this purpose.

Both Mr. Oliver and Mr. Fleming question my implicit

* L. G. Bostwick: *Bell System Technical Journal*, 1929, vol. 8, p. 135.
† *General Electric Review*, 1930, vol. 33, p. 505.
‡ *Experimental Wireless*, 1930, vol. 7, pp. 248 and 301.
§ *Journal of the Acoustical Society of America*, Oct. 1930.
‖ *Bell System Technical Journal*, 1931, vol. 10, p. 46.
¶ S. Hill: *Electrical Communication*, 1927, vol. 6, p. 24.

* H. Stenzel: *Elektrische Nachrichten-Technik*, 1930, vol. 7, p. 87.
† *Loc. cit.*
‡ Cf. the last words of Section 1(*a*) on page 77, and the first words of Section (11) on page 93.

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assumption that the mechanical impedance of the cone is itself unaltered by the presence or absence of acoustic loading. This is a perfectly sound point, and my own mind is open on the matter. It can only be settled by further work and comparison with direct measurements. I myself believe that the assumption is justified.

Mr. Hallawell suggests the use of a horn loud-speaker for the first experiment, as the measurement would have been easier. I agree completely, for the reasons he suggests. Unfortunately there was not one available.

Next, there are various points raised, not against the method, but in criticism of the particular results obtained in my example, or of my detailed methods.

Dr. McLachlan and Mr. Sowter imagine that I took no account of temperature effects on the measured resistance, but actually this was not the case. As I stated in presenting the paper, the bridge used could be arranged for d.c. resistance measurement by simply throwing a switch, and the d.c. resistance of the coil was checked from time to time. Mr. Sowter's estimate of the polarizing wattage was not a good one: the actual figure was approximately 9 watts. The loud-speaker reached temperature equilibrium in the period (about half an hour) allowed at the beginning of each series of tests for the source, etc., to warm up, and after this there was no appreciable variation of the d.c. resistance during the whole of the tests.

Mr. Sowter and Mr. Scroggie both apparently consider the method of " fixing " the diaphragm (loading with sand) to have been adopted as a makeshift or trouble-saver. This was not so. I was fully aware of the extreme difficulty of rigidly clamping a coil, and deliberately decided that it was better to increase the mass 100-fold than to try a rigid clamp. I was satisfied with the result when I found that the speaker was practically silent (even under large input) when so loaded: I still consider this to be a valid proof that the coil was in fact only moving to a negligible extent compared with normal. Mr. Sowter's point as to the reactance between 35 and 100 cycles does not impress me, and I fail to understand his remark as to fixing the coil " in the same position as for the free test "—the coil has no fixed position during the free test.

Mr. Sowter's remarks as to the " limited range " of my vibration galvanometer and on measurements with telephones at 150 cycles surprise me, as the Campbell vibration galvanometer has a normal range of 80–1 000 cycles. As to telephone measurements, anyone with a fair experience of a.c. bridge work soon learns to detect fundamental balance through harmonics.

Mr. Fleming points out some definite inaccuracies in my computations, for which I am most grateful to him: I have inserted the necessary corrections for the *Journal*. Luckily none of them affects the main argument.

Mr. Hallawell and Mr. Scroggie question my statement that the standing waves in the room gave rise to no perceptible changes of input impedance. I can only repeat that the point was foreseen and tested, as described in the paper, and no appreciable effect found.

Mr. Pocock's communication is interesting and sound, and I agree with him that some of the conclusions drawn from the particular tests described must be accepted with reserve, or, as I stated in the paper, are " speculative." But I must dispute his statement that I " base a good deal on the calculated air impedance for a disc in a baffle." I found that the acoustic reactance at low frequencies agreed well with the theoretical value (page 88) and that the acoustic resistance did not (page 89), and I made no further use of the theoretical figures. I stated that the curves of Fig. 1 were approximate only. I am very grateful for the reference to Stenzel's very important paper, which I had not previously seen.

Lastly, there is that very large proportion of the discussion which appears to me to be almost entirely irrelevant, dealing as it does with theories of loud-speaker action. In replying, I propose as far as possible to refrain from airing my own views on loud-speaker design, as being outside the scope of the paper.

I will deal first with the contributions by Dr. McLachlan and his assistant, Mr. Sowter, and first of all with

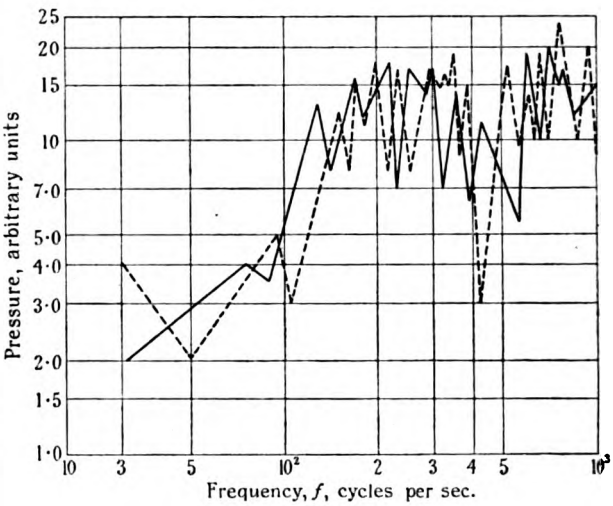

FIG. H.

the last point raised by them—the " conspicuous " absence from the Bibliography of reference to Dr. McLachlan's work. It should have been obvious that a " bibliography " of only three entries was not intended as a guide to the literature of the subject. It simply gives the references for publications which were in my mind as I wrote the paper. I am fully aware that Dr. McLachlan has written much on the subject of moving-coil speakers, and has been working on them for a long time. My only concern is to claim that if he had made, say, half of the 10 000 observations he mentions *in vacuo* as I now suggest, he would have learned even more than he has. As far as I can gather from their remarks, he and other contributors direct the most vigorous of their criticism to the fact that the computed response curve (Fig. 18) does not show the large number of sharp peaks and dips of the normal measured curve. I would first emphasize that this is fully explained by the procedure adopted, and should have occasioned no surprise. But there remains the important question: Are these peaks and dips in ordinary response curves " real " in the sense of appertaining to the speaker itself, and not to accidents of the surroundings? It has now

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for a long time been my view that most of these irregularities are quite meaningless from this point of view.

Since the reading of the paper, Mr. P. G. A. H. Voigt, by courtesy of Edison-Bell, Ltd., has very kindly taken a set of response curves for me on automatic apparatus in their laboratory, which records continuously from 32 to 8 000 cycles. Unfortunately, the original loud-speaker was not available, and the tests had to be done on a similar but not identical instrument. Curves were taken at 5 ft. distance at various angles to the axis, also on the axis at various distances. The curves are relative only, the microphone not being calibrated in absolute pressure. All the curves show the typical sharp peaks and dips; but every change in microphone position shows a complete redistribution of them. As an example, I show in Fig. H parts of two curves carefully transcribed from the original records to show all peaks and dips from 32 to 1 000 cycles—I have drawn straight lines from maximum to minimum points. One of these curves is for 3 ft. 9 in. and the other for 5 ft., both on the axis. This example and a careful comparison of the other curves on the same lines, confirm me in my opinion that these sharp irregularities are due to standing waves or similar accidents extraneous to the speaker itself, and that they are quite misleading in any attempt to judge its performance from the curve.

Dr. McLachlan and other contributors consider that the measurements must be in error because the speaker tested shows a steadily falling curve and not a large peak at 3 000 cycles or thereabouts. This again should have been expected. This peak at some high frequency is in many cases fictitious, due to measuring only at some point on the axis. If my computed curve means anything at all, it expresses integrated spherical radiation; and this 3 000-cycle hump is not then by any means a matter of course. As a matter of interest I have drawn curves from Mr. Voigt's measurements which omit those " fortuitous " peaks and dips which differ entirely on the curves for neighbouring microphone positions, and retain only such as appear several times. I then, for each of several frequencies (including all the remaining maxima and minima), summed the series of curves at equal distance and varying angle, taking the product of (pressure)^2 by (area of zone of sphere), and took the square root of the sum, thus getting the effective pressure for the whole of the radiated power. This result I corrected to constant voltage (for transformer losses) and for effect of finite baffle. The result is shown in Fig. J, with the computed constant-pressure curve of Fig. 18 dotted in for comparison. It is obvious that the smoothing procedure adopted masked the dip at 2 000 cycles and the humps at 1 000 and 4 000 cycles, but the general run of the curves shows good agreement: the humps are nearly, but not quite, within the limits I suggested in the paper (6 decibels). It remains to ask whether the method of measurement is capable of showing such dips and humps; and from Mr. Fleming's remarks, based on computations from the unsmoothed data, I am of opinion that it could. I much regret that I have no opportunity of taking check measurements at the higher frequencies, to confirm this.

It is of interest, as a matter of loud-speaker theory, to note from the curves taken at different angles that, although the tendency to " beam radiation " is quite obvious at 1 000 cycles and of course even more so at 3 000 cycles and above, yet the radiation at 2 000 cycles, where there is an important dip, is almost exactly spherical. This seems to me to disagree completely with any " resonance " theories, and to indicate peculiarities of radiation from a hollow cone (as distinct from a disc) as a more likely main cause of the dips and humps.

The remainder of Dr. McLachlan's contribution is an interesting statement of some parts of his theory of loud-speaker action, but has in general little to do with my paper. I desire to comment on two points. (1) Dr. McLachlan states that he heard " my " loud-speaker at the works of Messrs. Graham Amplion, Ltd. He does not state which of many experimental models he heard; contrary to his belief, it was not of the same design as the one tested. (2) He states in regard to one of his hypotheses " . . . there should be no need of experimental verification . . ." As far as my own experience of loud-speaker work goes, this seems to me to be a very misleading statement. The whole theory is still in the air, and it all needs checking by measurement.

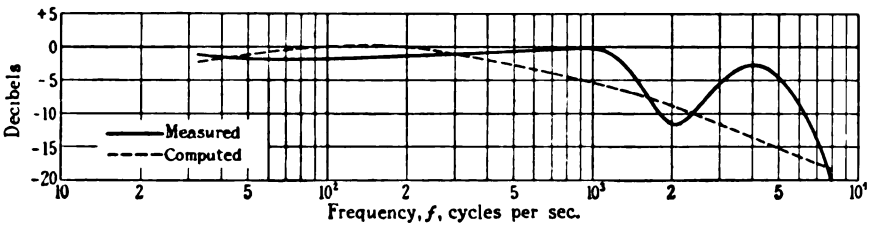

FIG. J.

Various contributors have read into my paper a statement I did not make, viz. that the cone can be fully represented by a simple transmission line. On page 83, par. 5 (*c*) (ii), I say *a* " transmission line " (not " simple "), and further on I expressly state that it cannot be fully represented by a simple line, and give in detail just how far I draw inferences on that basis. These inferences are guarded ones, and even so are not used in the subsequent argument except on page 85, par. 6 (*c*), where again the results are only given as " plausible." I am fully aware of the complexity of vibration of a cone: where I differ from these contributors is that I doubt whether this complexity has such important effects as they apparently expect, either on the acoustic output or the input impedance. I do not propose to go further into the matter here, as it is beyond the scope of the original paper. My own view has long been that most

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of the effects usually ascribed to what are loosely called " resonances " of the cone—a very vague term—are due to (1) accidents of the surrounding space, such as standing waves or other reflection effects, or (2) peculiarities of radiation from a hollow cone, as distinct from a piston. Stenzel's results, quoted by Mr. Pocock, compared with Fig. J, are very suggestive on this point.

Mr. Warren regrets the meagreness of my " basic analysis " and its " limited application to the problem in hand," but I feel he has failed to grasp what *is* the problem in hand. I gave the bare essentials required to convert the observed electrical impedances to their mechanical and/or acoustic equivalents: the problem in hand is this, not the theory of loud-speaker action. I do not recognize his term " frequency recurrence." If he means " an impedance showing periodic change with frequency," as seems likely from the context, then I dispute his assertions on the matter, which are offered without theoretical proof or results of test.

Summing up, the discussion has failed to shake my belief that the method of measurement described is a sound and powerful one, though it obviously requires more work before the results can be unreservedly accepted—as, indeed, was stated in the paper itself.

Finally, I should like to thank those contributors who have shown interest in the real point of my paper—the method itself. I must also state, with regret, that I cannot claim priority. Since writing the paper I have seen in the *General Electric Review* that the method was independently thought of by the General Electric Co. of America and has been adopted in their laboratories.

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THE PRACTICAL CORRECTION OF A WIRELESS DIRECTION-FINDER FOR DEVIATIONS DUE TO THE METAL WORK OF A SHIP.*

By C. E. HORTON, M.A.

(*Paper received 7th January, and read before the* WIRELESS SECTION *4th March*, 1931.)

SUMMARY.

The paper deals with the experimental development of direction-finders carried out at H.M. Signal School, Portsmouth.

A general treatment is given of the field of a wireless wave in the vicinity of a ship. It is shown that both semicircular and quadrantal correctors will generally be required to obtain readings of high precision and to correct a direction-finder for deviations due to the ship.

Typical forms of quadrantal and semicircular correctors are described and their relative advantages indicated.

The principles underlying a new form of sense-finder are explained.

The best position for a direction-finder in a ship is discussed and the results of measurements made in one of H.M. ships are given.

INTRODUCTION.

A wireless direction-finder is liable to give inaccurate results from a number of causes. It may be that the instrument itself, owing to imperfect screening, or to unsuitable circuits, is inaccurate even when used under ideal conditions with plane-polarized electromagnetic waves having the electric force vertical. Except at very high frequencies, however, such errors need not amount to 0·5° if suitable and well-known precautions are taken.

Again, errors are sometimes due to the influence of the ground or the ionization of the atmosphere in modifying the direct propagation of the wave from source to receiver. Such errors may be greater with some types of direction-finder than with others, but in general they cannot be entirely eliminated and are beyond complete control.

Lastly, errors may be due to the immediate surroundings of the direction-finder. A special case of this is a direction-finder in a ship, and it is the object of this paper to describe methods of correcting this error which have proved successful and convenient in practice.

The discussion of this problem will inevitably call to mind the correction of a magnetic compass for errors due to the magnetism of the ship. This is done by introducing forces which exactly counteract the effect of the ship's magnetism, that due to the hard iron of the ship being counteracted by permanent magnets, and that due to the soft iron by soft-iron correctors. Similarly, one may correct for the currents which flow in closed conducting paths in the ship by a loop suitably orientated and placed near the direction-finder, and for the displacement currents in the dielectric by a subsidiary aerial coupled inductively or otherwise to the direction-finder. But it should be noticed that there is an important difference between correcting for the steady magnetic field due to a ship's magnetism and for the alternating magnetic field set up by the action of a wireless wave on the conducting structures in the ship. In the first case there is always a precise direction for the resultant field which arises from the combination of the earth's magnetic field with that due to the magnetism of the ship; that is to say, the deviation is always a definite and measurable quantity. In the wireless case, on the other hand, the resultant alternating magnetic field will in general be elliptical and cannot therefore be said strictly to have a direction.

The degree of ellipticity of the field may be large or small, depending on the frequency and the direction from which the wave strikes the ship. If it is small the direction of the major axis of the ellipse may sufficiently define the direction of the wave, but if it is large the operation of taking a bearing with an ordinary direction-finder may become impossible or very uncertain and the term " deviation " is meaningless.

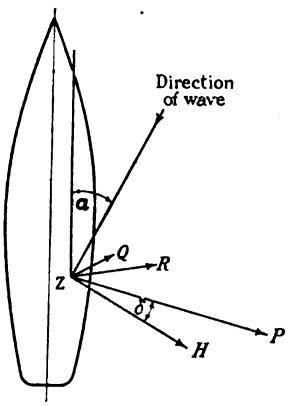

FIG. 1.—General case of a direction-finder at a position Z.

GENERAL ANALYSIS OF THE FIELD.

In order to examine this matter in greater detail consider a plane wave which strikes a ship from a direction making an angle α with the centre line. Alternating currents are induced in the hull of the ship, in the metal work of the superstructure, and in all aerials and other conductors.

At any point Z (Fig. 1), where we may imagine a direction-finder to be situated, each of these currents produces a subsidiary magnetic field. These fields will in general be different in magnitude, direction and phase, and their horizontal components may be denoted by

$$h_1 \sin(\omega t - \theta_1),\ h_2 \sin(\omega t - \theta_2),\ \text{etc.}$$

acting at angles β_1, β_2, etc., to the centre line, the alter-

* Reprinted from *Journal I.E.E.*, 1931, vol. 69, p. 623.

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nating magnetic field of the wave being represented by $H \sin \omega t$.

These components can be resolved into

$$h_1 \cos \beta_1 \sin (\omega t - \theta_1), \text{ etc.}$$

in the fore-and-aft direction and

$$h_1 \sin \beta_1 \sin (\omega t - \theta_1), \text{ etc.}$$

athwartships.

We may now proceed to combine together all the fore-and-aft components into a single resultant $Y \sin (\omega t - \eta)$, where Y and η are functions of

$$h_1 h_2 \ldots \beta_1 \beta_2 \ldots \theta_1 \theta_2 \ldots$$

That this is possible is easily seen by noting that two alternating forces

$$H_1 \sin \omega t \quad \text{and} \quad H_2 \sin (\omega t - \theta)$$

acting in the same direction are equivalent to a single force

$$H_0 \sin (\omega t - \gamma)$$

where $$\tan \gamma = \frac{H_2 \sin \theta}{H_1 + H_2 \cos \theta}$$

and $$H_0^2 = H_1^2 + H_2^2 + 2H_1 H_2 \cos \theta$$

In a similar manner we may combine the athwartship components into a single resultant $X \sin (\omega t - \nu)$.

Now resolve $Y \sin (\omega t - \eta)$ into two components $F \sin \omega t$ and $F' \cos \omega t$, both fore and aft but in quadrature with one another.

This means that $F = Y \cos \eta$ and $F' = -Y \sin \eta$.

Similarly resolve $X \sin (\omega t - \nu)$ into two components

$$B \sin \omega t \quad \text{and} \quad B' \cos \omega t$$

both athwartships but in quadrature with one another.

$F \sin \omega t$ and $B \sin \omega t$ may now be combined into a single resultant $R \sin \omega t$ at an angle

$$\epsilon = \text{arc tan} \frac{B}{F}$$

with the centre line, while $F' \cos \omega t$ and $B' \cos \omega t$ combine into a resultant $Q \cos \omega t$ at an angle

$$\psi = \text{arc tan} \frac{B'}{F'}$$

with the centre line.

The whole of the forces acting at the point Z are now reduced to:—

(1) A component $H \sin \omega t$ due directly to the wave, and directed at an angle $\alpha + 90°$ to the fore-and-aft line,
(2) A component $R \sin \omega t$ at an angle ϵ, and
(3) A component $Q \cos \omega t$ at an angle ψ.

$R \sin \omega t$ may be further combined with $H \sin \omega t$ to give a single resultant $P \sin \omega t$ and we are left finally with two fields, one $P \sin \omega t$ in phase with the wave acting at an angle ϕ and the other $Q \cos \omega t$ in quadrature acting at an angle ψ.

In general these two forces are in different directions and a rotating field is produced. A coil situated at Z has an e.m.f. induced in it no matter how it is directed. In other words, there is no zero and we have the familiar phenomenon of a blurred minimum. But if the coil is placed perpendicular to $P \sin \omega t$ only the quadrature component $Q \cos \omega t$ induces an e.m.f. in it.

Suppose now that a source of e.m.f., also in quadrature with the force of the wave, is available and that its amplitude can be made equal to $-Q$; then it is clearly possible to produce a perfect zero by injecting it suitably into the circuit of the frame coil. Furthermore, if the subsidiary e.m.f. is strictly in quadrature with the wave there is only one direction of the coil for which a zero is possible, namely when it is perpendicular to P. In any other direction the e.m.f. induced into the coil contains a component in phase with the wave, which cannot be balanced by any subsidiary e.m.f. in quadrature with the wave.

Thus a rotating coil and an adjustable source of quadrature e.m.f. may be used to give the direction of P and the magnitude of $Q \cos (\phi - \psi)$. In a similar manner with an adjustable source of e.m.f. in phase with the wave we may find the direction of Q and the magnitude of $P \cos (\phi - \psi)$. Hence both P and Q and the angle between them are determined.

The angle between the direction of P and the direction of H will be called the deviation δ.

Quadrantal Correctors.

So far we have dealt with the general case of a point Z in any position in the ship and have made no assumptions as to the actual magnitudes and phases of the subsidiary fields h_1, h_2, etc. In practice it is usual for Z to be situated on the centre line and this introduces a measure of symmetry into the forces involved.

Under very favourable circumstances the component Q is found to be zero, or very nearly so, for all directions of the wave. Other things being equal, this is more likely to happen with a small ship than with a large one, and usually demands a direction-finder supported in a high position clear of metal work. The effect of the ship is approximately the same as that of two conducting loops, one in a vertical plane through the centre line and the other in a vertical plane athwartships. For a wave incident at an angle α to the centre line, R consists of two terms $b \sin \alpha$ directed fore and aft and $f \cos \alpha$ athwartships, b being the component due to the beam loop and f that due to the fore-and-aft loop, and therefore b/f is small. The total force P acts at an angle

$$\text{arc tan} \left\{ -\frac{H + f}{H + b} \cot \alpha \right\}$$

with the centre line, giving an apparent direction of arrival of the wave of

$$\text{arc tan} \left\{ \frac{H + b}{H + f} \tan \alpha \right\}$$

which is nearer to the fore-and-aft line than α.

The deviation δ is zero for $\alpha = 0°$, $90°$, $180°$ and $270°$ and changes sign every 90 degrees. It is therefore termed quadrantal, and methods of correcting for it

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have been employed since the earliest days of direction-finders in ships.

The principal methods in use at the present time can be summarized as follows:—

(1) *Bellini-Tosi or Fixed Aerial Systems.*

(*a*) One loop is set up fore and aft and the other loop athwartships. The relative dimensions of the loops are adjusted by trial and error until it is found that a signal incident at an angle of 45° gives equal currents in the two aerials. To secure this result the fore-and-aft loop has to be made smaller than the beam loop. The ratio of the areas of the two loops depends on the size of the ship and on the position of the aerials with respect to the hull and superstructure. A ratio varying between 0·9 and 0·7 is common in battleships, while a ratio as low as 0·5 has been used with satisfactory results.

But it is clearly an advantage to have the ratio as near to unity as possible and it is of interest to note that in the vicinity of funnels and other large masses of metal this is best achieved by making the beam aerial as large as circumstances permit, and adjusting the fore-and-aft aerial accordingly. It is for this reason that large loops are preferred to small frames in such positions as between funnels.

An adjustment of aerials carried out on a wavelength of 1 100 metres is found to hold good for all longer waves. For shorter waves, however, the adjustment may not be correct. In a battleship the correct adjustment varies noticeably with the wavelength for waves less than about five times the length of the ship. Fig. 2, for example, shows the variation in δ with wavelength for a direction-finder in a ship of length 760 ft. and beam 88 ft. Between 1 200 metres and 450 metres the maximum deviation increases from 10° to 18°. Correction of quadrantal error by adjustment of aerials is therefore limited in practice to the longer waves.

(*b*) Starting with aerials of convenient dimensions, the equality of currents for a wave incident at 45° may be achieved by adding impedance in the circuit of the aerial which is receiving best. With the usual untuned aerial circuit, resistance is unsuitable because the amount required will vary with the frequency. To add inductance in circuit with each side of the loop is the usual practice, and the adjustment holds good so long as δ is constant, i.e. for the longer waves. For the shorter waves the adjustment varies with the wavelength and has to be ascertained by trial. With the receiving apparatus near to the aerials this method is a very satisfactory one, but in naval ships it is more usual for the instruments to be at a considerable distance (say 150 ft.) from the aerials. Account has then to be taken of the capacitance in the cables which join the aerials to the goniometer. This has the effect of making the relative impedances of the two aerial circuits dependent on the frequency.

Take, for example, a system with two aerials of equal inductance 100 μH joined by cables 150 ft. long to a goniometer the field windings of which have an inductance of 50 μH. With the best cables in common use at present the capacitance will be about 0·0008 μF. To correct a deviation of 10° on 2 000 metres will need a series inductance of 66 μH. At 800 metres this amount of series inductance will only suffice to correct 7·5°, and on 600 metres 3·5°. At still shorter wavelengths the correction changes sign. To correct 10° on 600 metres would require 210 μH, and still more inductance for shorter waves. This is unsatisfactory because the addition of series inductance tends to increase the natural wavelength of the circuit, and a point is soon reached at which the currents in the two field windings of the goniometer become appreciably out of phase.

(*c*) An alternative method is to shunt the field winding of the fore-and-aft aerial by inductance. To correct 10° on 2 000 metres needs a shunt inductance of 78 μH and the shunting becomes more powerful the shorter the wave, which is in the right direction because δ generally increases as the wavelength decreases. This is shown in Fig. 3, where it is seen that 78 μH will correct 13° on 600 metres and 16° on 450 metres.

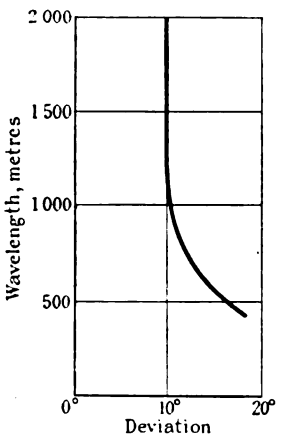

FIG. 2.—Variation of δ with wavelength.

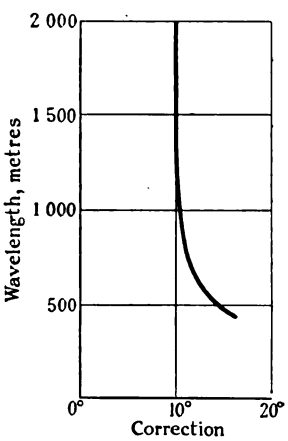

FIG. 3.—Correction of quadrantal error by shunt inductance.

Moreover, the addition of a shunt corrector tends to lower the natural wavelength of the circuit and does not therefore give rise to de-phasing of the aerial currents. It is of interest to note that by adding a capacitance of 0·00015 μF in parallel with the cable capacitance a single shunt inductance of 78 μH will correct δ for all waves down to 450 metres in accordance with the curve of Fig. 2, with an accuracy of 1°. The possibility of this automatic correction is an important advantage of this method. Another advantage is that a defective contact in the shunt circuit introduces an error at most equal to the maximum value of δ; whereas a defective contact in the series inductance method (*b*) may lead to a bearing up to 90° in error.

(2) *Rotating-Coil Systems.*

(*a*) A fixed loop surrounding the coil is set up in the fore-and-aft direction. The hull of the ship tends to increase the athwartship component of the flux through the coil. The currents induced in the loop tend to reduce the flux athwartship. With a suitable degree of coupling between the auxiliary loop and the rotating coil the flux through the coil can be adjusted to have the value appropriate to the undisturbed field of the wave. This method is chiefly of theoretical interest and is not easy to adjust for different wavelengths.

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(*b*) The difference between the true bearing and the bearing as given by a direction-finder is found in the first instance by trial.

With the ship at a distance of a few wavelengths from a transmitting station direct comparison is made between the wireless bearing and the true visual bearing for different angles of incidence of the wave relative to the fore-and-aft line. In this way a curve is obtained for the deviation. It is clear that by engraving a special scale it is possible to arrange that the direction-finder gives correct bearings directly. But the method has two disabilities. One is that the curve is only accurate over a limited range of wavelengths. For some services this may not matter and then the simplicity of the method has much to recommend it. A second disability arises when it is desired to obtain from the direction-finder true geographical bearings, and not merely

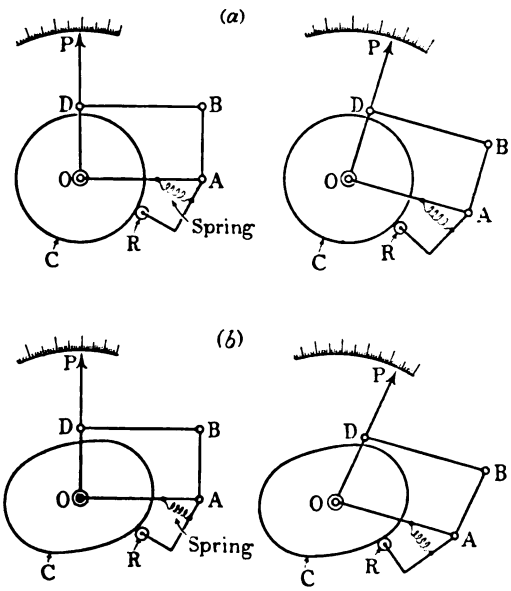

FIG. 4.—Diagram illustrating cam corrector.
(*a*) Roller bearing on circular section of cam.
(*b*) Roller bearing on elliptical section of cam and giving a correction.

bearings relative to the ship's head. To do this it is the common practice in H.M. ships to connect the scale of the instrument with the gyro-compass of the ship so that in effect the scale remains stationary in space. A scale specially engraved relative to the fore-and-aft line and with unequal divisions cannot be used in this way. It necessitates a separate gyro-compass repeater; true bearings are then obtained by adding the relative bearing given by the direction-finder to the bearing of the ship's head as given by the repeater. This operation takes time and gives the possibility of error.

It is always desirable to relieve the operator of the labour of computation. The method therefore adopted in mechanical forms of error correction is to operate on the pointer of the instrument so that it is displaced through an angle equal to the deviation.

A design of cam corrector developed in H.M. Signal School is shown diagrammatically in Fig. 4. As the coil is rotated by the handwheel (not shown) the arm OA moves always in step with it. The roller R is constrained by a spring to move on the surface of the cam C. AB, BD is a parallel link motion joined to the pointer OP and such that the angle between AR and AB is constant. Consequently for a circular cam there is no tendency for AR to rotate about the point A and therefore OP moves through the same angle as the coil. When, however, the cam is not circular AR revolves round the pivot A as the coil revolves. The angle between AR and AB being fixed, the rectangle OABD is distorted to a parallelogram and the pointer OP moves through an angle equal to the angle through which the coil has moved, plus the angle of rotation of the arm AR. In this manner an angular correction of any desired amount can be given to the pointer. When once the curve of deviation has been ascertained by trial it is only necessary to make a suitably-shaped cam to obtain correct bearings directly from the scale.

In many ships it is known before calibration what type of deviation will be encountered, and the cam is constructed in advance. Experience shows, and it has been deduced from elementary theory above, that δ is often of the form

$$\alpha - \text{arc tan}\left(\frac{H+b}{H+f}\tan\alpha\right)$$

For small values of f/H and b/H one may deduce that

$$\delta = \frac{1}{2H}(f-b)\sin 2\alpha$$

and the deviation is a maximum at 45°, 135°, 225° and 315°. To correct a deviation of this type by means of a cam the latter must be so shaped that the polar curve of the perimeter is of the form

$$r - a = k\sin 2\theta$$

θ being measured from a position of the roller where the correction is zero.

When f is not very small compared with H the maximum value of δ occurs when

$$\tan\alpha = \sqrt{\left[\frac{H+f}{H+b}\right]}$$

i.e. when the wave is incident at an angle of more than 45° to the fore-and-aft line. If a deviation of this type has a maximum amplitude of 10° it occurs when $\alpha = 50°$. For $\alpha = 45°$ the deviation is also very approximately 10°, but for $\alpha = 70°$ the deviation is 7·5°, while the simple type of quadrantal curve proportional to sin 2α would give a deviation of 6·5°. Similarly, for $\alpha = 30°$ the deviation is 8° while the curve proportional to sin 2α gives 8·7°. This has to be borne in mind in shaping a cam for corrections of 10° and more.

By constructing a solid cam it is possible to provide a range of values for the maximum quadrantal error. The cam illustrated in Fig. 4 is in fact a solid cam of depth $\frac{7}{8}$ in., and the particular cross-section of the cam on which the roller works can be adjusted by a screw in the centre carrying graduations 0 to 15. When the cam is so set that the reading is 0 the roller is bearing on a circular section of the cam near the face. In this position there is no correction. When the reading is 15 the roller

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is bearing on a section near the rear of the cam which is so shaped that the maximum correction is then 15°. Intermediate positions of the cam provide for maximum corrections having any value between 0 and 15°.

Although cam correctors have been used chiefly with rotating-coil direction-finders, it is clear that they can equally well be fitted to the pointer of a goniometer used with fixed aerials. When the deviation follows a simple law, electrical methods of correction are easily applied to fixed aerial systems and have advantages over mechanical correctors. When the law is very irregular. however, a cam corrector may be the only solution. An example of an irregular curve of deviation is shown in Fig. 5, which refers to a ship of length 760 ft. and beam 96 ft., carrying crossed loops 42 ft. to the starboard of the centre line, and 280 ft. from the bows. It will be noticed that there is a deviation of 3° for a wave incident from right ahead. It was established by trial in this ship that a similar aerial system located on the centre line gave no deviation right ahead. The off-centre position, however, was the only possible one

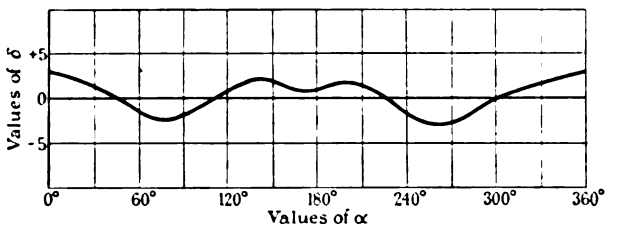

FIG. 5.—Example of an unsymmetrical curve of deviation.

for a permanent installation and its defects have been overcome by fitting a goniometer with a special cam corrector.

SEMICIRCULAR CORRECTORS.

Although a ship is usually built symmetrically about its centre line, it is never exactly symmetrical about a line drawn athwartships. In battleships the departure from symmetry is more noticeable than in most other ships. Similarly the ship's ordinary aerials are frequently symmetrical about the centre line but seldom about a line drawn athwartships through the mid-station of the ship. The result is that a direction-finder situated in any of the positions in which it is practicable to fit it suffers from blurred zeros. In other words, the component Q is appreciable.

Referring to Fig. 6, FA represents a coil on the centre line directed fore and aft, while PS represents a coil athwartships. Experience shows that when a transmitting station is exactly ahead or astern no signal is received by PS. But it is frequently found that the behaviour of the fore-and-aft coil FA is not so simple, and that when the transmitting station is exactly on the beam some signal is still received by the coil; in fact the coil is able to receive signals from every direction.

Thus for a station on the beam a rotating coil gives no real zero, but only a more or less ill-defined minimum. At the same time for a station right ahead or astern the zero may be perfect. The same results are, of course, obtained with a Bellini-Tosi set.

Blurred zeros on the beam are very common in ship direction-finders. Sometimes the blurring is slight and gives little trouble; sometimes it may very gravely reduce the accuracy with which a bearing can be taken. The higher the frequency the more marked the effect, and it is worthy of notice that a single loop PS with its plane athwartships can be used as a direction-finder, provided it is permissible to swing the ship through the bearing of zero signal. A good bearing can frequently

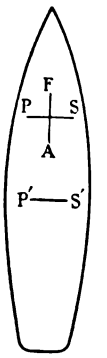

FIG. 6.

be obtained by this method at frequencies too high for an ordinary direction-finder to give a sharp reading in any direction except ahead or astern. But the method is generally too inconvenient to the ship to be practicable.

If an examination is made of the e.m.f. induced in the coil FA when the transmitter is on the beam, it is found in many instances to be in quadrature with the e.m.f. induced in the coil PS. This quadrature e.m.f. is due to re-radiation from the ship and its aerials, and it is clearly possible to balance it out and obtain a perfect

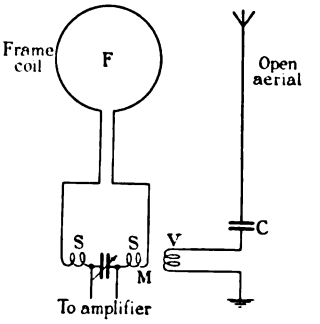

FIG. 7.—Circuit for open aerial corrector.

zero. All that is needed is a subsidiary source of e.m.f. in the right phase which can be injected into the circuit of the coil to oppose the e.m.f. due to the ship.

Such an e.m.f. may be obtained in a variety of ways. For example, a vertical aerial may be connected in circuit as shown in Fig. 7. Then, neglecting for a moment the fact that a small proportion of the energy picked up by the aerial will be due to re-radiation from the ship, it is seen that the e.m.f. put into the coil circuit from the vertical aerial will be in the right phase, provided only that the aerial circuit is far from resonance. This is most readily ensured by using a fixed series condenser C, for then it is easy to get an accuracy of phase of 1° over a very wide range of wavelength.

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If the e.m.f. in the coil due to the direct action of the wave is $A \cos \omega t$, then the e.m.f. due to re-radiation is $B \sin \omega t$ and the subsidiary e.m.f. from the vertical aerial $V \sin \omega t$. By means of a variable mutual inductance M it is possible to make $B = -V$ and then a perfect zero is obtained for a transmitter on the beam, exactly as if the ship were not there.

For a transmitter right ahead or astern experiment shows that $B = 0$ and therefore M should be zero for these directions in order to obtain silence when the receiving coil is athwartships. The coefficient B is a semicircular function of the direction α of the incoming wave. To a first approximation, and sometimes very closely, B is proportional to sin α. With a rotating-coil direction-finder and a semicircular corrector of this type, the correct coupling M, to obtain a perfect zero, therefore varies for different directions. The adjustment is simple in practice, the movement of the coil and the variation of the coupling resembling the

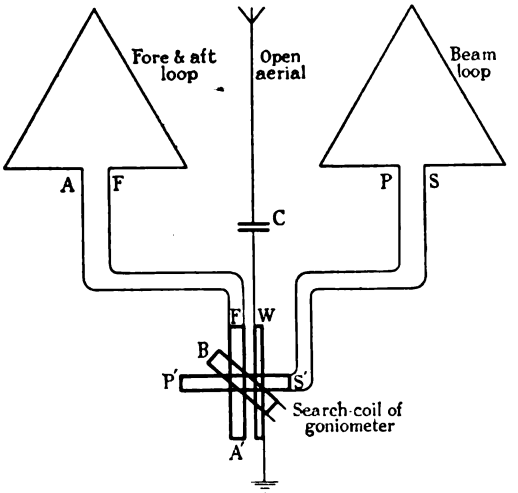

FIG. 8.—Fixed aerial system with open aerial corrector and special goniometer.

double-adjustment balance which is familiar in many alternating-current bridges. The adjustment could, of course, be made automatic by coupling the coil V mechanically with the frame F and so disposing the windings SS that the mutual inductance varies as the sine of the angular displacement of the coil V from the position it has when F is athwartships.

With crossed-loop direction-finders, however, a simple solution is possible and this is illustrated in Fig. 8. The vertical aerial is joined to earth through an additional winding WE of the goniometer, which is parallel to the winding F'A' connected with the fore-and-aft loop. It will be seen that for a wave incident right ahead the search coil B will be parallel to P'S' when a zero is obtained, the vertical aerial then contributing nothing; while for a wave incident on the beam the search coil will be parallel to FA, the vertical aerial then contributing sufficient to cancel the blurring due to re-radiation.

Instead of using an open aerial to obtain a subsidiary source of e.m.f., it is possible to use a closed loop P'S' (Fig. 6), sometimes with advantage. One circuit which gives the correct phasing is shown in Fig. 9, the coupling M between the two tuned circuits being small.

As the e.m.f. induced in a loop varies with the direction of the wave, it is best to set up the loop athwartships, without appreciable coupling to the direction-finder aerial itself. The induced e.m.f. is then zero for $\alpha = 0$ and 180° and a maximum for $\alpha = 90°$ and 270°, while for intermediate directions it is proportional to sin α. If B is also proportional to sin α it is clear that a balance

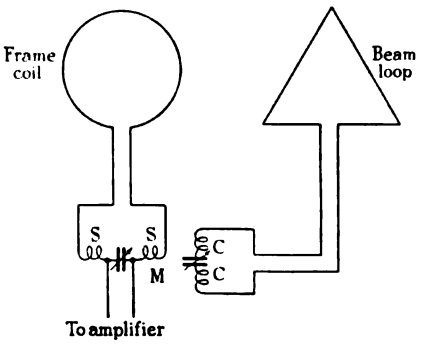

FIG. 9.—Circuit for beam loop corrector.

can be obtained for all values of α by a fixed coupling M. This is a valuable property of a subsidiary loop used with either a rotating coil or fixed aerials. But it will be noted that the correct functioning of the circuits depends upon precise tuning of one or more circuits. It is well known that to control phases in this way is a very delicate matter and offers difficulty to an operator; moreover, there is a very serious possibility of error due to the presence of subsidiary components in phase with the wave, tending to give a false bearing.

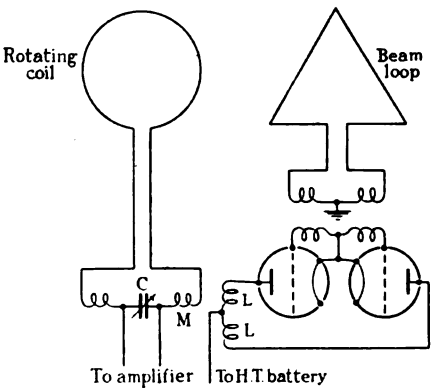

FIG. 10.—Alternative circuit for beam loop corrector.

For this reason the practice is to make use of a valve circuit as shown in Fig. 10. If the impedance of L is small compared with the impedance of the valve the current in the anode circuit is in quadrature with the voltage applied to the grid. To secure an accuracy of phase of 1° in this way demands an impedance for L not more than about 1 per cent of the impedance of the valve. When this is done, however, the correct phase relation between the coil e.m.f. and the subsidiary e.m.f. is maintained, even if the tuning condenser C is not exactly in resonance.

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Ambiguity of Bearing.

Picture a ship at anchor heading due north and carrying a direction-finder. Imagine that a second ship acting as transmitting station steams round on the arc of a circle from the position W due west of A to the position E due east of A.

Using a rotating-coil direction-finder and an open aerial corrector of the type shown in Fig. 7, bearings are taken of the transmitting ship as she steams round. It is found that the correct coupling M is a maximum with the ship at W and decreases to zero when the ship is at N due north of A. The coupling has then to be reversed and gradually increased in magnitude until the ship is at E. In this position (and similarly at W) there is only one direction of the coil for a perfect zero. If the coil is rotated through 180° from this position there is not a true zero but only a minimum of sound, the amplitude of which is proportional to $2B$. This minimum may, however, be converted to a true zero by reversing the coupling M. The combination of the rotating coil with this type of corrector is therefore not a sense-finder, although one zero is perfect and the other is not. To restrict the sign of M to positive values merely restricts the useful arc of the frame coil to 180°.

The corrector shown in Fig. 8 must necessarily give two opposite zeros which are always of equal quality. The subsidiary winding WE affects both zeros equally and therefore this arrangement also is not a sense-finder.

Consider now a beam loop corrector of the type shown in Fig. 9. It is found that the correct coupling M for the ship at W holds very closely as the ship steams to the position N, and continues to hold without reversal of the coupling as the ship steams to the position E. There is only one direction of the coil for a true zero. By fixing the sign of the coupling M the arrangement becomes a sense-finder having the very valuable property that the sense zero is in the same direction as the true bearing. For bearings exactly ahead or astern, however, both zeros are perfect and an ambiguity still exists.

This corresponds to the case of the perfect direction-finder on shore giving two sharp zeros 180° apart. In order to resolve the ambiguity it is usual to employ some form of receiving system with a cardioid characteristic. This has two disadvantages. In the first place the minimum of the cardioid is displaced 90° from the two original zeros and therefore necessitates a separate sense-pointer, and secondly the determination of sense is an additional operation which is done after the bearing has been taken accurately with the figure-of-eight characteristic. This entails a loss of time, necessitated by the fact that the cardioid itself is not generally suitable for taking accurate bearings; for it is obtained by opposing the main loop e.m.f. by a subsidiary e.m.f. obtained from an open aerial, and these have to be brought into an exact equality of phase and amplitude in order that the bearing may be correct. Such a balance is difficult to maintain with the precision required to get correct bearings in all directions.

To overcome these defects a sense-finder has been developed by the author, in conjunction with Mr. Cecil Crampton; B.Sc., also of H.M. Signal School, which has the merit of blurring the reciprocal zero while leaving the true zero as sharp as before and in the original position. In other words, the operations of taking a bearing and determining the sense are one and the same, exactly as with the beam loop corrector described above. The device is illustrated in Fig. 11, in which F_1 is the main rotating coil and F_2 an auxiliary coil at right angles to F_1 and rotating with it. V is an open aerial not coupled directly with F_1 or F_2. The coil F_1 alone gives two zeros 180° apart. When the aerial V is added both zeros are blurred, but by the further addition of the coil F_2 one of these blurred zeros can be converted into a perfect zero while the other becomes still more blurred.

We may express the action symbolically by considering a wave incident at an angle α from true north. The

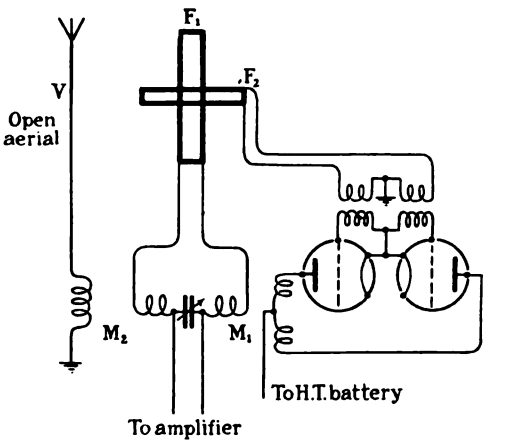

Fig. 11.

e.m.f. induced in the receiver when F_1 is at an angle θ is of the form

$$f_1 \cos(\alpha - \theta) \sin \omega t + f_2 \sin(\alpha - \theta) \cos \omega t + v \cos \omega t$$

It is therefore necessary for a zero that

$$f_1 \cos(\alpha - \theta) = 0$$

and

$$f_2 \sin(\alpha - \theta) + v = 0$$

If f_2 and v are both positive this implies that

$$\theta = \alpha + 90° \text{ and } f_2 = v$$

This is the only value of θ for a zero, the reciprocal bearing $\theta = \alpha - 90°$ giving a finite e.m.f. $2v \cos \omega t$. Similarly, when the direction of the wave is reversed the zero is only obtained when $\theta = \alpha - 90°$. There is therefore no ambiguity.

It will be observed that the whole system amounts to the addition of a cardioid to a figure-of-eight, with the following conditions:—

(1) The e.m.f. of the cardioid is in quadrature with the e.m.f. of the figure-of-eight.
(2) The zero of the cardioid coincides in direction with one zero of the figure-of-eight.

The actual amplitude of the cardioid is theoretically immaterial, provided the phasing is correct. But the

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essential advantage of the scheme is that the sense is completely determined without prejudice to the accuracy of the bearing when the amplitude of the cardioid is only a small fraction of the amplitude of the figure-of-eight. An imperfection in the cardioid either of shape or phase gives only a negligible error of bearing.

If, for example, the phasing is correct but f_2 and v, though small, are not exactly equal, then neither zero is quite perfect, but one is sharper than the other. Nevertheless, the position of minimum sound is still when

$$\theta = \alpha + 90°$$

exactly as for the perfect zero.

Again, suppose that there is a slight error of phasing of the auxiliary loop, so that the total e.m.f. is now

$$f_1 \cos(\alpha-\theta) \sin \omega t + f_2 \sin(\alpha-\theta) \cos(\omega t-\epsilon) + v \cos \omega t$$

where ϵ is small and f_2 and v both small compared with f_1. A zero is obtained when

$$f_1 \cos(\alpha - \theta) + \epsilon f_2 \sin(\alpha - \theta) = 0$$

and

$$f_2 \sin(\alpha - \theta) + v = 0$$

whence

$$\cos(\alpha - \theta) = \epsilon v / f_1$$

Thus θ differs from $\alpha + 90°$ by a small quantity of the second order. It is, for example, amply sufficient for the determination of sense if v is 5 per cent of f_1, and then an error of phasing of 5° causes an error of bearing of 0·25°.

A sense-finder similar in operation to the beam loop corrector is thus available for circumstances in which no suitable auxiliary e.m.f. is present due to the action of surrounding structures. It is obvious that by using a goniometer with a search coil carrying two windings at right angles the same result can be obtained with a Bellini-Tosi system. It is also of interest to note that the principle of this sense-finder can be applied to the transmitting beacon to give one direction having a true zero of radiation and a direction 180° from this with a finite amount of radiation.

Effect of the Ship on the Correcting Aerials.

It has been assumed so far that the field acting on the subsidiary aerial or loop has the same phase as the undisturbed field of the wave. This cannot, of course, be quite true; though it may be nearly true for an aerial high above the deck.

If the electric force in the wave is $E \sin \omega t$, the electric force in the vicinity of the correcting aerial can be represented as $\lambda \sin \omega t + \mu \cos \omega t$, where μ is usually small compared with λ. It has been shown that the total field acting on the direction-finder can be reduced to $P \sin \omega t$ at an angle ϕ to the centre line, and $Q \cos \omega t$ at an angle ψ to the centre line, with Q usually small compared with P.

The e.m.f. induced in a coil set at an angle θ to the centre line is proportional to

$$k_1\{P \cos \omega t \sin(\theta - \phi) - Q \sin \omega t \sin(\theta - \psi)\}$$

and the e.m.f. induced in the coil circuit from the vertical aerial is proportional to

$$k_2\{\lambda \sin \omega t + \mu \cos \omega t\}$$

A zero can therefore be obtained provided

$$k_1 P \sin(\theta - \phi) + k_2 \mu = 0$$

$$k_1 Q \sin(\theta - \psi) - k_2 \lambda = 0$$

whence

$$\tan \theta = \frac{P\lambda \sin \phi + Q\mu \sin \psi}{P\lambda \cos \phi + Q\mu \cos \psi}$$

The position of the coil for a zero is therefore perfectly definite and unique and cannot be influenced by the operator, who can only rotate the coil and adjust the ratio k_1/k_2. When Q/P and μ/λ are small the difference between θ and ϕ will be a small quantity of the second order and the error obtained by neglecting μ is very small. Consider, for example, a direction-finder situated in a position where the in-phase component R is sufficient to give a quadrantal error of 5°, while there is a constant quadrature field Q directed athwartships of amplitude about one-tenth of the total in-phase field P. Assume also $\mu/\lambda = 1/10$.

Then it is found that the maximum value of $\theta - \phi$ is 0·5°. When $Q = \frac{1}{4}P$ and $\mu = \frac{1}{4}\lambda$ the maximum value of $\theta - \phi$ is 3·5°, and, allowing for the fact that P varies in amplitude by about 20 per cent for different

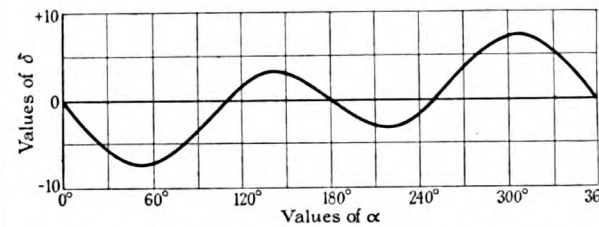

Fig. 12.—Theoretical curve of deviation for $Q = \frac{1}{4}P$.

directions of the wave, it is found that the deviation is as shown in Fig. 12.

It will be observed that δ is zero when $\alpha = 0$, 110°, 180° and 250°, and that the deviation is no longer quadrantal.

With a quadrature component as strong as this it would be very difficult to take a bearing by finding the position of minimum sound without using a correcting aerial; moreover, the position of minimum sound in this case, even if it could be ascertained, is not exactly the same as the position of the true zero which is obtained by the use of the correcting aerial. For example, with a wave incident at 45° the true zero is obtained when $\theta = 127°$ if a correcting aerial is used. This corresponds to an apparent bearing of 37°. The minimum obtained with the simple rotating coil occurs when $\theta = 132°$, corresponding to an apparent bearing of 42°. For small quadrature fields δ does not depend appreciably on the position of the subsidiary aerial, but for larger values of Q/P, as may be obtained with short waves, the direction-finder must be calibrated and used with a definite aerial in a definite position.

It has been said that in many instances Q is directed athwartships, so that a beam loop corrector does not pick up an appreciable quadrature component. The position of the coil for a zero is therefore always at right angles to P and in the particular case considered above,

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and using this type of corrector, the curve of deviation crosses the axis at 90° and 270° and is truly quadrantal.

Stated in general terms, the problem of making a direction-finder which will give true bearings directly by scale readings can be resolved theoretically into getting two e.m.f.'s, E_1, E_2, one of which (E_1) is in phase with the wave, and the other (E_2) in quadrature, and injecting E_1, E_2 into the circuit so as to counteract exactly the distortion of the wave due to the ship. E_2 eliminates blurring and E_1 corrects the deviation. Generally E_1 has four zeros in the 360°, and E_2 two zeros. Under favourable circumstances there is symmetry in the disposition of these zeros and automatic correctors can be provided by simple means. Sometimes a symmetry is observed but not about the usual axis. In one ship, for example, with aerials aloft, but off the centre line, the zero of E_2 occurs for $\alpha = 165°$ and 345°. A loop exactly athwartships is then unsuitable for automatic correction in accordance with Fig. 10, but a loop inclined 15° to the beam, i.e. from 75° to 255° relative to the centre line, gives a satisfactory solution.

In the most general case, however, E_1, E_2 do not follow precisely any simple law and it is necessary to find by trial exactly how they vary for different values of α. To make the correction automatic it is then necessary to couple the rotating coil mechanically with two variable mutual inductances whose law of variation is adjusted by means of specially constructed cams to give the correct values of E_1, E_2 for any position of the coil. At present it is the practice to have only one cam and to adjust E_2 separately by hand. It should be noted that E_2 need not be exactly adjusted for a perfect zero in order to get the correct bearing. A slight blurring may be unimportant. But where time permits of the perfect zero being obtained the precision of a single reading can be very high—better than 0·5° with continuous waves. Under such circumstances the greater the amplification the more sharply defined the zero becomes.

Best Position for a Direction-Finder in a Ship.

It is not generally possible to choose the site for a direction-finder merely from electrical considerations, but experience gained in many ships has shown what is to be avoided and what can be achieved. The obvious position for a direction-finder is as high above the hull of the ship as possible. Apart from any difficulties in transferring the received voltages from the aerials to the office, experience confirms this. The quadrantal deviation becomes less with increasing height. One of the most sensitive tests of a good direction-finder, however, is the quality of the readings it gives. The exact amount of the deviation is of less importance, provided it is known and constant. In other words, it is desirable to put the direction-finder where Q is small. For although Q may be balanced by an auxiliary e.m.f. it is an advantage from every point of view to have it small. If Q is considerable on long waves it may become the preponderating term at short wavelengths, and bearings will be difficult to obtain. There is no hope for a direction-finder on short wavelengths except in a position where Q is very small on long wavelengths.

Referring to Fig. 13, a direction-finder of the screened rotating-coil type is shown at A on top of the fire-control position of a cruiser. The coil is 3 ft. 6 in. in diameter and is rotated from an office at a distance by apparatus designed by the Admiralty Research Laboratory. This apparatus is of great value in making it possible to put rotating coils and goniometers in positions normally inaccessible but favourable to direction-finders. Moreover, it has been pointed out to the author by Dr. Drysdale, Director of Scientific Research to the Admiralty, that quadrantal errors may be corrected by adjustments to the electrical elements of the apparatus. In this way the coil can be caused to lag or lead on the pointer in the office so that the latter gives true bearings without correction.

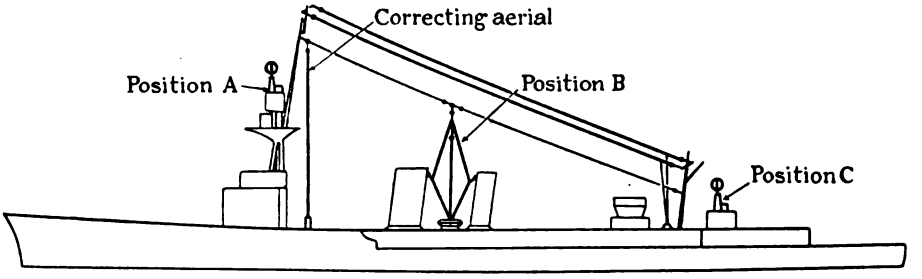

Fig. 13.

It would appear at first sight that a position so high above the deck as that shown at A would be a very favourable one. But it has to be remembered that the wooden mast immediately behind the coil carries many earthed conductors such as lead-cased cables for mast-head lights, main aerial halyards, lightning conductor, etc. All these contribute subsidiary fields in their immediate vicinity, and experience shows that the zeros obtained on either beam are badly blurred, and render the direction-finder of little value unless a form of quadrature corrector is also fitted. This may, for example, be a single-wire aerial as shown. Readings of high precision can then be obtained in all directions. Fig. 14 shows the curve of deviation for a wavelength of 2 700 metres, obtained by experiment under conditions of weather favourable to accuracy. This curve holds within 1° down to 1 100 metres, the maximum deviation being 3·5°. At 450 metres the curve takes the form shown in Fig. 14 by the dotted line, crossing the axis at 104° and 255°, which is in accordance with the theory

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based on Q being directed athwartships and approximately constant for different values of α.

The magnitude of the quadrature e.m.f. necessary to obtain a perfect zero can be found by the use of apparatus such as is employed in measuring field intensities. The procedure is as follows. With a wave incident at a definite angle α a true zero is obtained by rotating the coil to the position ϕ and adjusting the mutual inductance M. The coil is now disconnected, and the e.m.f. induced in the receiver through the mutual inductance M is measured. This gives Q, and, of course, varies with

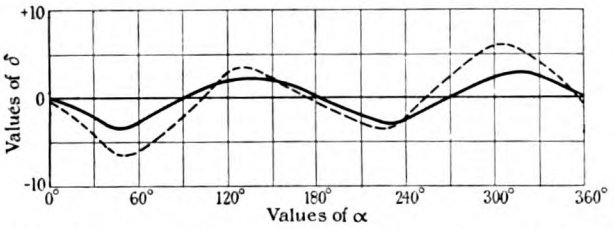

FIG. 14.—Variation of δ with α for wavelengths above 1 100 metres and for a wavelength of 450 metres.

the intensity of the incident wave. But Q/P is independent of the strength of the signal and an approximate measure of P is readily obtained by setting the coil in a direction $\phi + 90°$, disconnecting the subsidiary aerial and measuring the e.m.f. induced in the receiver. The operation is repeated for different values of α from 0 to 360°.

For position A in the ship shown in Fig. 13 the values of Q/P found by experiment are shown graphically in Fig. 15. It will be seen that the curves are semicircular in type. On 2 700 metres the maximum value of Q/P is 0·05, on 1 100 metres it is 0·12 and on 450 metres 0·22.

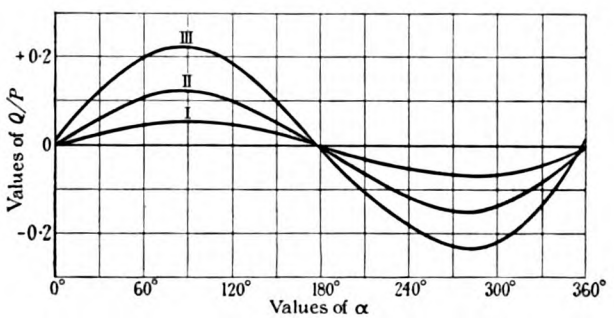

FIG. 15.—Variation of Q/P with α, (I) for 2 700 metres, (II) for 1 100 metres and (III) for 450 metres.

With quadrature components of this magnitude the position is a poor one for a direction-finder.

In some ships the construction is such that it is possible to put the coil on top of the mast and above the level of other aerials. A very great improvement is then found in the performance. Quadrature components on the beam are, however, still found, due to the action of the ship as a whole on the wave, and it appears that these may be greater for wavelengths of the same order as the length of the ship than for very short waves of, say, 20 metres.

A rotating-coil direction-finder situated in the position C abaft the mainmast gives a maximum deviation for the longer waves of 11° with quadrature components intermediate between those obtained for the two positions A and B. This position for a direction-finder is also comparatively poor without quadrature correctors.

A Bellini-Tosi system with large aerials fitted between the funnels as shown at B gave Q equal to 1 per cent of P at 2 700 metres, 2 per cent at 1 100 metres and 4 per cent at 450 metres. P and Q are here mean values throughout the space occupied by the aerials, and would not be the same for small frames placed between the funnels. This represents a comparatively good direction-finder requiring only small quadrature e.m.f.'s to give perfect readings. It is a common experience that the best direction-finders are those in ships where it has been possible to fit them on the centre line approximately amidships, with good symmetry both fore and aft and athwartships. A position immediately to one side of a single funnel is much inferior to a position between two equal vertical funnels.

The position of the main and other aerials relative to the direction-finder is generally an important factor, for it is a great advantage, and sometimes a necessity, to be able to receive on one aerial in the ship on exactly the same wavelength as that on which the direction-finder is working. This can be done with a rotating coil at the top of a mast and above the level of other aerials. But for a direction-finder under the main aerial it is found that Q varies with the tuning of the aerial. It is generally least when the aerial is insulated, and comparatively small when the aerial is connected to earth but far from resonance. In the immediate vicinity of resonance it is usually impossible to take a bearing, owing to the fact that the aerial gives strong subsidiary components both in phase and in quadrature. The effect of the aerial is the more pronounced as the direction-finder approaches the end near the feeder, and this position is usually avoided.

NIGHT EFFECT.

A point of interest in the use of quadrature correctors is their behaviour during periods of "night effect." For example, in the ship shown in Fig. 13 and using an open aerial corrector, the bearing of 5XX during a period of night effect immediately after sunset varied between 346° and 9° from North, the true bearing being 359°. The necessary coupling M to obtain a true zero varied at the same time between two limits proportional to sin 7° and sin 82°, the normal daytime value for this bearing being proportional to sin 49°.

The daytime values of M for different bearings are generally known fairly closely after the set has been in use a little time. Under conditions of night effect, when a portion of the energy reaches the ship from the upper atmosphere, the distortion produced by the ship is different from that produced on a wave travelling horizontally. The normal relation of amplitude and phase between the e.m.f. induced in the direction-finder and that induced in the vertical aerial is destroyed.

When, therefore, it is found necessary to employ unusual values for M in order to get a zero, night effect is at once suspected. In this way the corrector is of considerable value in telling the operator when every-

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thing is behaving normally, and has frequently led to the detection of sources of error which might otherwise have passed unnoticed. With a beam loop corrector properly adjusted a clear zero is obtained automatically for all directions of the wave, provided the direct ray is alone being received. If, therefore, a blurred zero is obtained at any time in the subsequent use of the apparatus, it is an indication of the presence of a downcoming or atmospheric ray, and the apparent bearing must be viewed with suspicion.

This paper is published with the permission of the Board of Admiralty. The author desires to thank the Captain, H.M. Signal School, Portsmouth, for assistance in the preparation of the paper. In addition, he would like to express his indebtedness to Mr. G. Shearing, B.Sc., Member, Director of the Experimental Division; also to his colleagues of the Civilian and Naval Experimental Staff of H.M. Signal School for assistance in the design of apparatus and in the organization and conduct of trials at sea.

Discussion before the Wireless Section, 4th March, 1931.

Mr. G. Shearing: There are probably no applications of direction-finding where the complexity is greater than it is in naval ships, employing as they do a very wide range of frequencies and having large masses of metal in the superstructure. The author's comparatively simple solution of some of these problems is one of the outstanding features of the paper. It is important to note that the semicircular or quadrantal-error corrector (Figs. 7 and 8), the beam loop corrector (Fig. 10), and the combined sense-finding and direction-finding system (Fig. 11) each have only one tuning adjustment, and even if the operator does not attain best adjustments of the coupling required for the corrector devices he cannot, under any circumstances, interfere with the phase conditions. He cannot, therefore, obtain a bearing which is false as regards minimum signal strength, even though with poor adjustments the minimum-signal position may still be somewhat blurred. Thus the apparatus is virtually foolproof. This simplicity of operation is an essential feature for the successful design of modern naval wireless apparatus. Under the heading "Ambiguity of Bearing" on page 113 the author describes a non-ambiguous circuit. This circuit developed by the author and Mr. Crampton is important, since it is a true radio-compass system and will undoubtedly find considerable application in the future, probably not only to direction-finding but also to transmitting beacon systems, as is pointed out in the paper. In connection with the theory of this circuit, it is interesting to note that the variation of signal intensity is proportional to the square root of the sum of the squares of the intensities due to the figure-of-eight diagram produced by the rotating coil (F_1) of Fig. 11, and the quadrantal cardioid produced by the open aerial (V) in conjunction with the auxiliary loop F_2. I should be glad if the author could indicate this in a diagram. The essential point is, however, that although the additional signal strength obtained for a given position of F_1 does not follow any arithmetical law, the single zero is virtually as clear and definite as would be obtained for the two zeros of the figure-of-eight diagram. The statement on page 115 that there is no hope for a direction-finder on the short wavelengths except in a position where the equivalent re-radiated component Q is very small on long wavelengths, epitomizes the requirements for the direction-finding loops in a ship. The difficulty as regards short waves is, of course, due to the fact that at the higher frequencies the metal work of ships' rigging may correspond to the resonating conditions, and so increase the component Q upon which the ellipticity of the field depends. There is experimental evidence that the equivalent component Q is greatest for wavelengths which correspond dimensionally to the actual length of the ship in which the bearings are being taken. With regard to the Bellini-Tosi system referred to on page 109, it is interesting to note the author's ingenious method of correction for deviation of the fore-and-aft loop by using a shunt inductance across the field winding, in conjunction with the capacitance of the fairly long feeder connected to the goniometer in the office, in order to obtain automatically a constant correction over a wide range of frequencies. The problems connected with the fitting of direction-finding apparatus in naval ships bristle with difficulties. In many cases one has to put up with what position is left after all other requirements in the ship have been met, and the paper shows how some of the problems have been satisfactorily solved.

Mr. S. B. Smith: The practical correction of a ship direction-finding installation presents many interesting problems, particularly when the wavelength range extends from 400 to 4 000 metres and the accuracy must be within $\pm 2°$. In the introduction to the paper it is stated that errors due to propagation phenomena commonly known as "night effect" cannot be entirely eliminated and are beyond complete control. This statement does not truly represent recent advances made in the art, as it is now quite possible to produce land direction-finders capable of reducing these errors to a negligible order, and in general such installations will provide a practical aid to navigation even during periods of severe night effect. The elimination of night errors in a marine direction-finding installation, however, presents additional problems. On land, where the aerial system can be isolated from masses of metal and the principles laid down by Adcock can be applied, a commercially accurate night-effect-free direction-finder is already available. On board ship, however, large masses of metal are known to have directional receiving properties somewhat similar in character to those of a loop aerial, and consequently re-radiation may impair the efficiency of night-effect-free aerial systems installed upon a vessel. I should like to ask the author whether any experiments have been carried out on board ship with a view to ascertaining the magnitude of night errors when a shore type of aerial system is installed. Considerable research work seems to be necessary before it will be true to say that night-effect-free ship direction-

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finders are commercially practicable. In mercantile marine navigation the ranges and operating conditions are such that no demand appears to exist for an accurate night-effect-free direction-finder, but for naval purposes such an instrument would seem to possess distinct advantages. Fig. 2 is interesting and clearly indicates how restricted the wavelength range should be for a given calibration when working between 500 and 1 000 metres, in the particular case shown. Has the author attempted to extend this curve to cover the short-wave band? On page 109 is given a typical example of the constants of a Bellini-Tosi loop system connected by means of paper cables to the radiogoniometer, and it is seen that the natural wavelength of the system is approximately 300 metres. What is the lowest wavelength on which crisp and accurate bearings can be obtained with this arrangement? Regarding the correction of quadrantal errors in the Bellini-Tosi system, it is pointed out in the paper that a suitable calibration is obtained by altering the ratio of the areas of the two loops until it is found that a signal incident at an angle of 45° gives equal currents in the two aerials. Although in certain cases this method is successful, an overall swinging of the ship is still desirable to ensure that the maximum uncorrected error is negligible. When installing Bellini-Tosi direction-finders it is common practice (where possible) to swing the vessel uncalibrated, and, having determined the nature of the error curve, to insert a shunt calibrating choking coil across the appropriate aerial. In rare cases it is necessary to rotate the aerials with reference to the centre line of the vessel, if the physical and electrical axes are not coincident. The present practice of employing small multi-turn aerials instead of large single-turn loops introduces the following instrumental difficulties:—(*a*) Sharp and well-defined minima, difficult to correct. (*b*) The calibration is more dependent upon masses of metal in the immediate vicinity of the aerial. (*c*) The signal/noise ratio is inferior to that experienced with large aerials, as in Fig. 13, position B. Commercial considerations, however, have led to the standardization of small loops in the mercantile marine. One of the many advantages of the Bellini-Tosi system as developed by the Marconi Co. is that greater freedom of choice of location is possible than with the rotating-loop method. By experimentally settling upon a suitable location (this position may be a considerable distance from the direction-finder), it is possible to improve the signal/noise ratio, and in many cases the overall calibration is more accurate. I observe that remote control of loops and goniometer is being employed, and I would ask the author whether this method has any advantages over the existing types of Bellini-Tosi direction-finders. In Fig. 13, position A, is shown an elevated loop direction-finder and it is stated that unless some form of quadrature correction is employed all the minima are blurred. The vertical reception on the cables connecting the loop to the receiver would appear to be very pronounced in such cases, and in my opinion the poor definition of the minima is chiefly due to this cause rather than to masses of metal in the immediate vicinity of the loop. Has the author made certain that the feeding cables from the loop to the receiver are incapable of producing the bulk of these quadrature effects? In connection with quadrature correctors, it is interesting to note that this method has now been in use on certain commercial direction-finders for some years. In practice the signal/noise ratio is often poor, and the method of taking zero bearings in conjunction with a minima clearing device possesses grave disadvantages as compared with small-swing bearings on a Bellini-Tosi apparatus. Under ideal conditions, with no time limitations in taking bearings, unskilled operators and zero noise-level, the zero method of working would seem to be attractive. In commercial Bellini-Tosi direction-finders the added complication of minima clearing devices has not been found essential to accuracy. The method of eliminating ambiguity of bearing shown in Fig. 11 seems rather complicated when long wavelength ranges are to be employed. The same operating advantages can be obtained by using two search coils in a Bellini-Tosi radiogoniometer and so arranging the circuits that one search coil is employed for taking the bearing, while the other is used in conjunction with an open aerial for obtaining the sense indication. The cardioid methods are extremely simple and rapid when applied to a radiogoniometer; for heavy rotating frames the author's proposal has distinct operating advantages.

Mr. R. H. Barfield: I am not clear how often the variable coupling in Figs. 7, 8 and 9 has to be adjusted in order to make the quadrantal component of the right amplitude. Has the operator to alter it for every bearing he takes, or can it be fixed once and for all? I can appreciate that, as the author states, the variation in the direction of arrival of the wave will be automatically compensated for as regards this coupling, but I should like to know whether or not there will be any " creeping " of the latter, due, say, to masses of metal being moved about on the ship, or to other causes. I am particularly interested in the author's analysis of the phenomenon commonly found in direction-finders of one minimum being flat and the other sharp. I had already concluded that it must be due to at least three separate and more or less independent e.m.f.'s, but the explanation has until now never been entirely clear. I think that the author and his colleagues are to be congratulated upon turning a drawback into an advantage by using this effect as a sense-finding device. Mr. Smith appears inclined to think that the quadrantal component appearing in the author's analysis is virtually the same as " vertical effect." I cannot see, however, that there can be any confusion between the two, for whereas the quadrantal component is always associated with the presence of local conductors, the phenomenon known as " vertical effect " may be present in a direction-finder even when it is erected in an open space.

Dr. R. L. Smith-Rose: The problem of making direction-finders give bearings to an accuracy of 0·5° on land is not an easy one, but to obtain this accuracy on a steel ship literally covered with metal wires, turrets and guns is an exceedingly difficult proposition. I think the author is to be congratulated upon having tried for so many years to solve the problem, apparently with a fair measure of success. His achievement is all the more remarkable in view of the fact that he is not

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likely to have had a free hand in choosing the position of the direction-finder in the ship. I gather that an accuracy of 1° or even less is being aimed at, and I sometimes wonder whether such a high accuracy is really necessary in a ship, despite the fact that the accuracy of navigation in the Royal Navy is probably on a higher plane than in the mercantile marine. I notice, for instance, that Fig. 5 shows a deviation curve in which the maximum error is 3°. Is it really necessary to reduce this to 0·5°? Is not 3° accurate enough for many purposes, and when greater accuracy is really required would it not suffice to make corrections from this error curve? Incidentally, I should like to know how accurately these correcting devices remain constant during a gale. Also, what order of accuracy is obtainable with the type of compass used in the ship? This question is important because it should be remembered that the bearings are taken relative to the direction of the ship's head, and therefore in order to translate these into geographical bearings one should know this direction to a higher degree of accuracy than is aimed at in the direction-finder. It was at one time considered that the single-loop direction-finder provided all the advantages which could be obtained by means of the Bellini-Tosi system, with its crossed coils and goniometer, except that on board ship the quadrantal error of the Bellini-Tosi system with large loops was claimed to be much smaller than that obtained with small rotating or fixed loops. It is therefore somewhat remarkable to find that those responsible for the development of the Bellini-Tosi system have of recent years employed quite small loops. They still retain the Bellini-Tosi principle and the accompanying goniometer, for reasons which appear justifiable, but I should like to know what has become of the excuse that was made some years ago for the use of large loops. Mr. Smith has raised the question of the possibility of eliminating " night effect " on direction-finders, and those of us who have tried the Adcock system have always been aware of the limitations with which we were faced in applying that system to direction-finding in ships. Usually, of course, sufficient space is not available to make the Adcock system sensitive enough to act as an efficient direction-finder.

Commander L. V. Morgan: In connection with the use of direction-finding apparatus in the Navy, I should like to emphasize, from my own experience at sea, the necessity of being able rapidly to obtain a quick bearing and foolproof sense in one operation. The circuits shown in Fig. 11 have an important application in this respect.

Mr. C. E. Horton (*in reply*): Mr. Shearing has drawn attention to a feature of the sense-finder described on page 113 which results from the fact that the e.m.f. due to the figure-of-eight characteristic is in quadrature with the e.m.f. due to the cardioid. Under these circumstances it is not legitimate to superpose a figure-of-eight on a cardioid and obtain the curve of total e.m.f. by direct addition.

In Fig. A the result of adding the small cardioid to the figure-of-eight, with due regard to phase difference, is shown by the dotted line. It will be seen that there is only one direction of zero response and 180° from this there is a minimum of finite amplitude equal to the amplitude of the cardioid.

Regarding Adcock direction-finders in ships, it may be stated in reply to Mr. Smith's question that such systems have been rigged experimentally in naval ships, but up to the present have not proved entirely satisfactory. To get accuracy of bearing the four aerials must have exactly equal receiving power, while for reasonable sensitivity they must be spaced an appreciable fraction (say, one-tenth) of a wavelength apart. These two requirements are difficult to satisfy in a ship on medium wavelengths.

The curve of Fig. 2 refers to a ship carrying large Bellini-Tosi aerials between the funnels. Such a system is not suitable for short wavelengths and the curve has therefore not been extended below 450 metres.

Regarding the Bellini-Tosi loop system given as a typical example on page 109, the lowest wavelength on which sharp bearings are obtainable depends on the position of the aerial loops in the ship, but is seldom less than 500 metres.

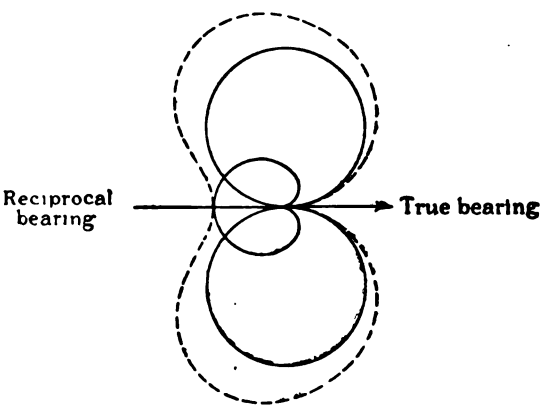

FIG. A.—Addition of a cardioid to a figure-of-eight in quadrature.

The remote control of rotating coils and goniometers has many advantages in naval ships, where the best positions for the aerials are often very inaccessible. Moreover, no difficulties arise due to cable capacitance causing resonance. In the ordinary Bellini-Tosi system this sets a distinct limit to the permissible length of cable. There is no such limit when the goniometer is remotely controlled.

Mr. Smith questions whether vertical reception on the cables between the aerial and the office is not the chief cause of blurred zeros for position A in Fig. 13. If this were so, blurred zeros would be obtained for all directions of the wave, whereas it is found that very sharp zeros are obtained without any quadrature correction for waves incident from right ahead or astern.

In reply to Mr. Barfield, it is true that with the open aerial type of quadrature corrector the correct coupling varies for each direction of the wave relative to the ship, so that it generally needs a fresh setting of the corrector for each bearing. With the beam loop type of corrector it is not necessary to vary the coupling after it has once been set correctly for one bearing.

Dr. Smith-Rose has raised the question of the degree

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of accuracy required in direction-finding. For tactical purposes, as emphasized by Commander Morgan, a rough bearing with absolute certainty of "sense" may often suffice. This is especially the case when the situation is changing rapidly and the true bearing is altering from minute to minute. On the other hand, an accuracy of $\pm$ 1° is very desirable for navigation. The gyro-compass repeater giving "ship's head" can be made accurate, with suitable precautions, to $\pm \frac{1}{2}°$, and therefore in general every effort is made to reduce instrumental defects in the direction-finder to the order of $\pm$ 1°. Only in this way can the probable error of the final result be brought down to a reasonable figure.

DISCUSSION ON

"A GRAPHICAL METHOD FOR DETERMINING THE MAGNITUDE AND PHASE OF THE ELECTRIC FIELD IN THE NEIGHBOURHOOD OF AN ANTENNA CARRYING A KNOWN DISTRIBUTION OF CURRENT." *

Mr. R. M. Wilmotte (*communicated*): Two years ago Mr. McPetrie and I co-operated in some research, and our only serious difference of opinion was whether certain quantities should be positive or negative. I am glad to see from this paper that he has successfully settled those questions. The paper should prove of value in the design of antenna systems, and I hope that the National Physical Laboratory will make the curves available in a more convenient form; by that I mean curves plotted to a larger scale on paper which has not had most of the co-ordinate lines removed. I want particularly to refer to the effect of the image in the ground. I am aware that the author has quoted me, but since the publication of the paper to which he refers I have to some extent changed my theory. Sommerfeld's equations show that the simple reflection theory holds asymptotically, i.e. only at large distances, and experiments confirm this to some extent. The extension of Sommerfeld's analysis to horizontally polarized rays seems to show that the simple reflection theory holds at quite small distances for these rays. I should like to make the following suggestion to explain the physics of this phenomenon. For vertically polarized rays Maxwell's equations and boundary conditions show that a ray travelling horizontally to a conducting surface must have a certain kind of elliptic polarization. The waves emitted from a vertical antenna are simply vertically polarized, and, in order to change this state of polarization into the steady elliptically polarized condition, currents are set up in the ground which produce a comparatively local field. It is this field which completely upsets the simple reflection theory when applied in the neighbourhood of the antenna. If the antenna were designed to radiate horizontally waves having the same polarization as that corresponding to a steady state of polarization, I believe it would be found that the simple reflection theory would hold even at small distances from the antenna. Such a design would have important practical applications because it would decrease the currents in the ground that are required to produce the change in the polarization of the ray propagated horizontally, and thus reduce the ground losses in the neighbourhood of the antenna. In the case of horizontally polarized rays, the boundary conditions show that they cannot be propagated horizontally. In that case there can be no ground currents to alter the field in the neighbourhood of the antenna and it is probable that the simple reflection theory will hold. It will probably be found for most practical conditions that in the case of vertically polarized rays a close approximation will be obtained for the field in the immediate neighbourhood of the antenna if it is assumed that the ground is a perfect conductor.

Mr. J. S. McPetrie (*in reply*): I am glad to see that Mr. Wilmotte has continued his interest in antenna systems since he left the National Physical Laboratory. The curves given in the paper are rather small for accurate computation; the original curves from which these were reduced have been made available to the British Post Office and any interested party could doubtless have copies on application to the Laboratory. Mr. Wilmotte's remarks on the effect of the earth are very interesting, but I suggest that his physical explanation is inaccurate. The field from a vertical dipole is elliptically polarized near the earth's surface, not because the incident wave is elliptical but because the earth currents set up by the incident wave with elliptical or plane polarization give rise to a reflected wave which is elliptically polarized. These earth currents produce a strong local field, which becomes sensibly plane at a small distance above the earth. It is only under this condition that Mr. Wilmotte's equations for the reflection coefficient can be applied, i.e. they cannot give the value of the field on the earth's surface but only at such distances above it that the wave reflected from the earth's surface is sensibly plane. The fact that horizontally polarized waves cannot be propagated along the earth's surface does not mean that there are no earth currents; on the contrary, the absence of propagation in this direction is due to the neutralizing effect of the re-radiation from the earth currents set up by the incident wave.

* Paper by Mr. J. S. McPetrie (see page 40).

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SHORT-DISTANCE OBSERVATIONS ON LONG-WAVE PHENOMENA.*

By R. NAISMITH, Associate Member.

(*Paper first received* 3rd *December*, 1930, *and in final form* 28th *January*, 1931; *read before the* WIRELESS SECTION 1st *April*, 1931.)

SUMMARY.

The experiments described show that, contrary to the usual assumption, reflection from the ionized layer plays a part in the propagation of long waves, even at relatively short distances.

The intensity of the reflected wave is found to depend very markedly upon the direction of propagation, even for an aerial which appears to have no favoured direction of ground-wave transmission. For transmission from south to north the reflected wave is found to be stronger than for transmission from east to west, which strongly suggests that the earth's magnetic field plays a part in the effect.

The results appear to indicate a correlation between short-distance (under 400 km) and long-distance (over 1 000 km) observations, which is not found to exist when correlation is attempted with observations over medium distances.

Previous experience on long-wave transmissions has shown that during the day-time in the summer the influence of down-coming waves from the upper atmosphere is only appreciable at distances of 300 km and upwards.

In some experiments carried out on the transmission from GBR (Rugby, $\lambda = 18\,700$ m) at Slough, which is only 111 km from Rugby, surprisingly large variations of intensity were found during the 6 months' observation period. Careful experiments showed that this effect was undoubtedly propagational in origin, and demonstrated that on occasion the magnitude of the downcoming wave might be 40 per cent of the direct or ground wave.

These striking results were further emphasized by the fact that signals were very different at two places nearly equidistant from the transmitter, although simple theory would lead one to expect close agreement.

POLAR CURVE OF GBR (RUGBY, $\lambda = 18\,700$ M).

In order to investigate this effect fully it was necessary to know the effective height of the aerial. It was also necessary to find out whether the aerial had a direction of maximum radiation. It was impossible to make accurate measurements of the effective height at Slough owing to the presence there of the reflected wave.

In experiments made at distances up to 45 km there was no evidence of a downcoming wave. At distances greater than 35 km the induction field is less than 6 per cent of the radiated field, but as it is in quadrature the effect may be neglected. The ideal zone for the measurements was therefore at 35 to 45 km from the station, and all effective-height measurements were made in this zone. The standard Hertzian formula

$$h = E_g d\lambda / 120\pi I$$

where h, d, and λ are in metres, E_g is in volts per metre, and I is in amperes, gives the theoretical values for comparison.

Observations were made all the way round the station except in the north-east quadrant. The lack of observations in this quadrant was due to a sudden change in the Rugby programme of transmission, which would have necessitated a prolongation of the tour. As the agreement round the other three quadrants was so definite there appeared to be no reason for supposing any appreciable departure from normal in the first. Fig. 1 shows the results obtained. If the Rugby aerial is assumed to be in the centre of the plan, the radii represent the direction in which the receiving site was situated from it. The length of any one radius represents the value of " effective height " calculated from the intensity measurement made at the site indicated by that particular radius.

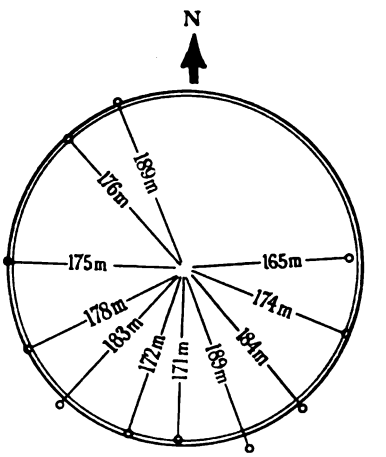

FIG. 1.—Rugby polar curve.
Frequency = 16 000.
Average effective height = 177 m.

From these results it would appear that (*a*) there is no direction of maximum radiation, and (*b*) the average effective height taken all the way round the station is 177 m.

This proof of symmetrical radiation from the aerial renders it possible to calculate the intensity of the direct wave from the above formula at distances up to 200 km from the aerial with a fair degree of accuracy. The medium through which the space wave travels must evidently be responsible for the surprising results described.

INTENSITY MEASUREMENTS AT MANCHESTER.

Observations made daily at Manchester over a period of several months had shown that the intensity of the

* Reprinted from *Journal I.E.E.*, 1931, vol. 69, p. 875.

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signals from Rugby was very much lower than the calculated value. For example, the average measured intensity at Manchester for the month of January 1927 was less than 8 mV per metre, whereas the calculated value for the direct wave was about 14 mV per metre. During this month the wireless bearing of the station also varied between the true bearing and 14° from the true bearing in a counter-clockwise direction. It is significant to note that the bearing never rotated clockwise, except occasionally during a magnetic storm.

In connection with the difference of signal intensity obtained at Manchester and Slough, it should be mentioned that these two stations are not at exactly the same distance from Rugby, and in order to obtain a perfectly valid comparison it was decided to make measurements at two points exactly equidistant in opposite directions from Rugby.

Simultaneous measurements were therefore made at Tadworth (Surrey) and at Manchester. The intensity of the signal at the former place was 18·6 mV per metre, while at the latter it was only 11·4 mV per metre. This striking difference clearly demonstrates the importance of direction of transmission when distances are identical. The result may, of course, be partly due to the different radiations upwards of the GBR aerial system, but the considerations below suggest that this is not the whole cause.

Measurements in Different Directions from Transmitting Stations.

A tour was arranged from Slough to Exeter during the summer, and measurements were made on the transmissions from Rugby and Northolt at each stop. Measurements were taken on Northolt to find out to what extent comparison could be made with the Rugby results. In the case of Rugby, there was no evidence of a direction of maximum intensity anywhere on the route taken during this tour, but the intensities measured were sometimes 50 per cent of the calculated values. Observations made at the same sites on the return journey varied considerably from the previous values, showing that the day-to-day variations at Slough were repeated elsewhere.

In the case of Northolt ($\lambda = 6\,950$ m) the bearing was approximately the same at all sites, but the distances varied from 50 to 250 km. The curve plotted in Fig. 2 shows these results. The smooth curve is drawn through the calculated intensities. The measured intensities all agree very closely with the calculated values over this distance, and there is no definite evidence of a reflected wave. The contrast is very striking, but the explanation may lie in the fact that the path of the signals from Northolt was always approximately east to west. In the case of Rugby the transmission was more nearly north to south.

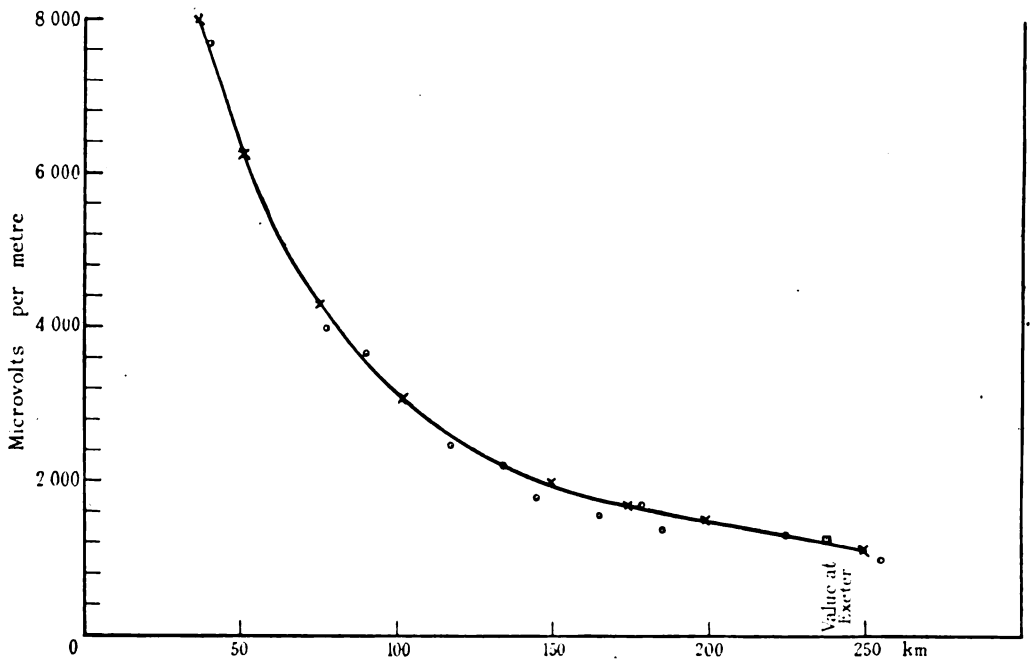

Fig. 2.—GKB ($\lambda = 6\,950$ m).
× Calculated values. ⊙ Measured values.

It was not possible to test this explanation by a tour on Rugby from east to west owing to the proximity of the Welsh hills, where it is always difficult to find suitable sites for measurement purposes. As measuring apparatus was already installed at Manchester and Exeter it was possible to test this interpretation further by choosing a suitable transmitting station. Northolt appeared to answer the requirements fairly well. It is almost equidistant from Manchester and Exeter. In the one case the wave is travelling approximately in the plane of the magnetic meridian, and in the other more nearly at right angles to it. It is readily admitted that the geographical conditions were not ideal, but the

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arrangement had the great advantage that the experiment could be started immediately, since the apparatus was already installed. It was therefore decided to try it out. Daily observations were made at these two stations for a period of 5 weeks, and the results are plotted in Fig. 3. From this it is apparent that at Manchester the average intensity was about 100 per cent greater than at Exeter. From the calculated value of the direct wave it is clear that the effect of the indirect wave was very much greater at Manchester than at Exeter. The bearings given at the foot of Fig. 3 are the calculated values, and in order to obtain the bearing with reference to the magnetic meridian 15° should be added to these figures. To test this conclusion fully, a totally different station working on a different wavelength was chosen. In this case the transmitting station was Ongar, working on 5 100 m. The receiving stations were Manchester and Exeter, as before, but the

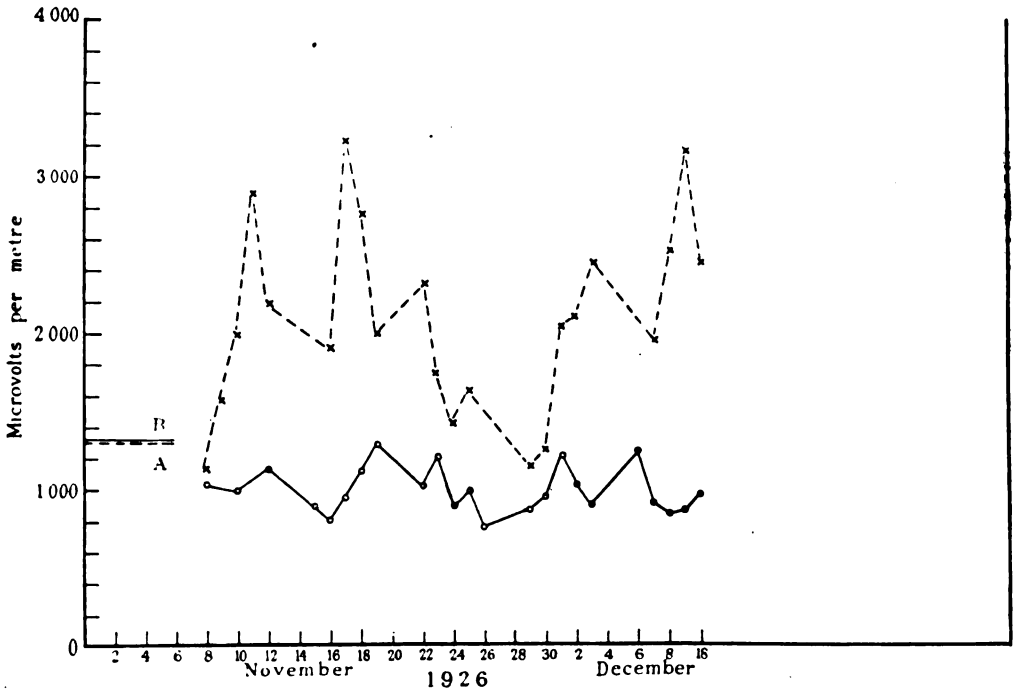

FIG. 3.—GKB (λ = 6 950 m).

- - - Manchester (distance 243 km, bearing 149). —— Exeter (distance 238 km, bearing 66).
A.—Austin-Cohen value (Manchester). B.—Austin-Cohen value (Exeter).

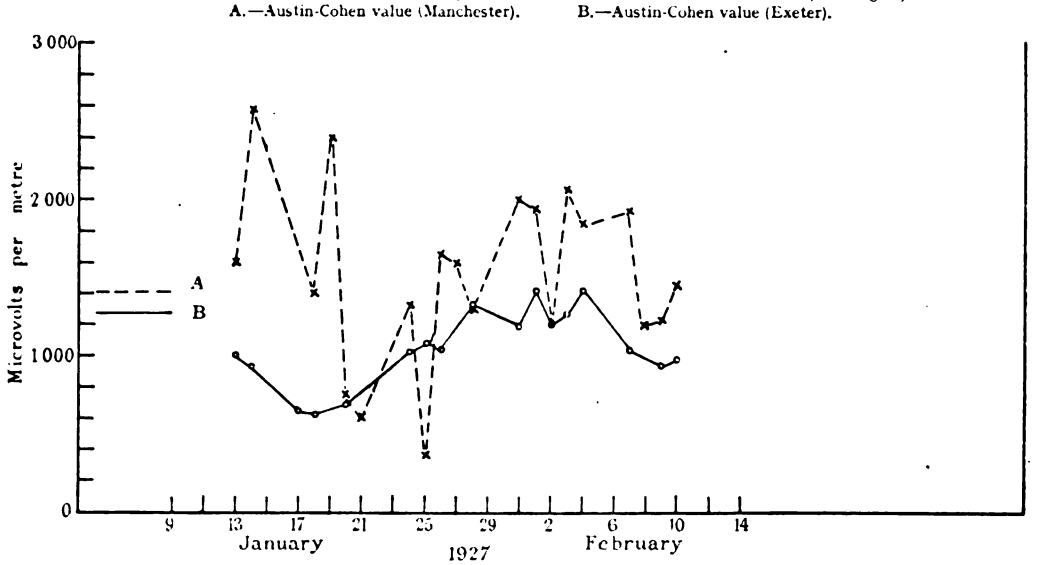

FIG. 4.—GLP (λ = 5 100 m).

- - - Manchester (distance 250 km, bearing 139). —— Exeter (distance 281 km, bearing 69).
A.—Austin-Cohen value (Manchester). B.—Austin-Cohen value (Exeter).

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distances were slightly greater. Daily observations were made for a further period of 4 weeks, and the results, which are plotted in Fig. 4, again show the same effects. Thus similar results are obtained from three different transmitting stations working on three different wavelengths. This justifies the conclusion that the medium through which the wave is propagated is responsible for the effects enumerated, and no further consideration need be given to the particular type of transmitting system in use. Further, since only a comparison between two readings taken simultaneously is required, the necessity for knowing the actual characteristics of the transmitter is avoided.

This particular directional effect had already been noticed during 1922 and 1923 by an expedition sent to Australia. "There was a period of several days between Wyndham and Colombo during which no signals were heard from the New York stations, this period corresponding with the time at which the great circle between the S.S. 'Boonah' and New York was over or near the North and South Poles."* It was suggested that the mixture of light and darkness over the path may have been the cause of the attenuation of the signals. Measurements made at New Southgate, England, on the transmissions from New York† lend no support to this theory. In fact, taking the monthly average curves of diurnal variation of field strength, in *every* case for 12 consecutive months the intensity between sunrise in England and sunrise in the United States (i.e. mixture of light and darkness) was actually higher than at any other period of the 24 hours, except when total darkness covered the path of the signal.

Comparison Between Short-Distance and Long-Distance Transmissions.

A considerable amount of work has been done by the Radio Research Board on medium-distance long-wave transmission (400 to 1 000 km), but no evidence of this directional effect has been obtained. The fact that a directional effect is observable on both short-distance and long-distance transmissions suggests a possible correlation between these two.

It must be remembered that the amount of energy leaving any transmitting station is definitely limited, and if conditions are such that a large amount of that energy is returned close to the transmitting station, there will be less to reach greater distances.

According to the paper referred to above,* measurements made in South Africa show that, metre-ampere for metre-ampere, the signal strength of the European stations to the north is only 50 per cent of the signal strength of the American stations to the west. The difference is largely attributed to land versus sea attenuation, though it is admitted that this effect could hardly account for such a big discrepancy. In the case of Rugby the path of the signals to the North Pole is by sea, yet one finds that the wave has suffered much more attenuation over this path than it has by traversing the all-land route to India.

It has already been shown that conditions are such that at short distances more energy is returned in the south-north than in the east-west direction, and therefore less energy is left to travel to greater distances in the south-north than in the east-west direction, which is in agreement with the results quoted.

Additional evidence of this connection between short-distance and long-distance transmission is afforded by T. Minohara,* who gives the results of a series of measurements made at the naval experimental and research establishment, Tokio, on signals from Pearl Harbour—a distance of 6 400 km. He compares the variations of intensity observed with changes of the prevailing atmospheric conditions, and arrives at the same conclusions as Dr. Austin, who obtained his results from measurements made over a distance of only 200 km. Further confirmation of this connection is obtained in magnetic-storm data. It had been found over a period of several months that, at short distances, magnetic storms did affect the transmission. This is in general agreement with experience of commercial working over long distances. In particular, on October 14th–15th, 1927, Eskdalemuir observatory reported the largest magnetic disturbance of the year, when the range of the north component reached 719γ, the west component 960γ, and the vertical component 636γ (1 gauss = 100 000γ). On that occasion the transmissions from Rugby were being measured at Manchester, a distance of 136 km. During the period referred to the intensity varied from 8·4 to 18 mV per metre, while the reflection coefficient rose to nearly 3 times its normal value. Comparing this effect over a short distance with measurements made over long distances and published in U.R.S.I. Document 127, the graphs of three long-wave stations for the two months September and October, 1926, are reproduced in Fig. 5. Particular reference to Rugby and Leafield shows that during the 11 days following this magnetic storm only one observation was made on the former station and two on the latter. The whole path of the signal was in daylight.

It has been noticed during the course of the observations made for this paper that the effect of the storm generally lasts for several days. It is therefore interesting to note that in U.R.S.I. Document 125 (November 1926) T. Watkins reports that on the 15th April, 1926, New Zealand experienced the most brilliant aurora for years, and that this was one of the worst nights for radio reception. Following this experience, it took 3 weeks for radio conditions to become normal again.

Magnetic storms, which are essentially night phenomena, reach their minimum intensity about midday. The observations recorded in this paper, with the exception of the Manchester results for October 14th–15th, were all made between 10 a.m. and 2 p.m. (G.M.T.) in order that the least possible disturbance from magnetic storms might be experienced.

Since the above experiment was recorded the radio laboratory of the U.S. Bureau of Standards has published an article† which concludes that at times of magnetic storms short-distance long-wave stations increased their intensity at Washington, while the intensity recorded

* H. J. Round, T. L. Eckersley, K. Tremellen and F. C. Lunnon: "Report on Measurements Made on Signal Strength at Great Distances During 1922 and 1923 by an Expedition Sent to Australia," *Journal I.E.E.*, 1925, vol. 63, p. 956.

† *Bell System Technical Journal*, 1925, vol. 4, p. 470.

* *Journal of the Institute of Electrical Engineers of Japan*, March 1927, No. 461, p. 225.

† *Science News-Letter*, 11th May, 1929.

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on long-distance stations was low. It must, of course, be noted that a mere statement of increased intensity is not unambiguous where two or more waves are being received, as in the case of short-distance long-wave transmission. It was a realization of this feature which caused the experiment incorporating the reflection-coefficient measurement to be performed at Manchester, as described above.

variation. This may be due to a gradual change in the height of the reflecting layer, with a consequent change in the phase relation of the direct and reflected waves, or it may be due to a gradual change in the polarization or reflection coefficient of the reflected wave. Typical sunset cycles are given in a paper by J. Hollingworth.*

Observations on the Rugby transmission were next made at Slough and Manchester over the usual sunset

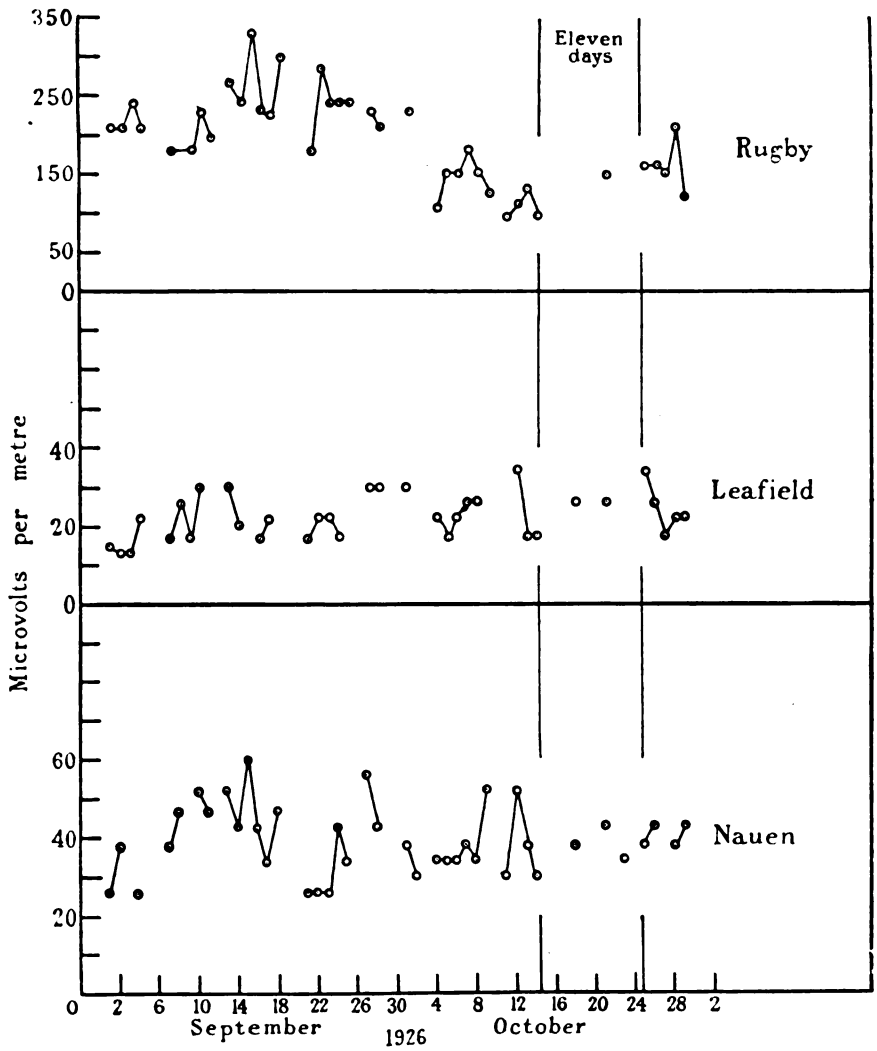

FIG. 5.

Comparisons made between short-distance and long-distance transmissions having shown such direct correlation on the simple energy argument, a comparison is now made between short-distance (less than 400 km) and medium-distance (400 to 1 000 km) transmissions.

COMPARISON BETWEEN SHORT-DISTANCE AND MEDIUM-DISTANCE TRANSMISSIONS.

It has been found that in the summer, for about 2 hours just after sunset, transmissions on wavelengths from 3 000 to 20 000 m at distances between 400 and 1 000 km go through a perfectly definite cycle of intensity

period, but the typical cycle of intensity variation was not obtained. In general, it was impossible at Slough to detect any change between these few hours of darkness and the day. That is to say, if there were no signs of a reflected wave during the day, the intensity remained constant during the 3 hours watched. If the reflected wave produced variations of intensity during the day, variations of the same order persisted during the period of sunset.

Sunset observations on the transmissions from Northolt were made at Manchester and Exeter during

* *Journal I.E.E.*, 1926, vol. 64, p. 579.

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the summer. These showed a variation at Exeter of less than 5 per cent, but the intensities recorded at Manchester varied erratically from 1·2 to 1·5 mV per metre during the period of observation. It should, of course, be remembered that even during the day-time in the summer, the intensity of Northolt's signals at Manchester varies considerably. This is as one might expect, for the height at which the waves are reflected, in the case of high-angle reflection, can hardly be expected to possess the same sharply defined limit as that of the lower ionized layer, which gives rise to interference effects of the type shown by Hollingworth to exist at the medium distances of 300 to 1 000 km from the transmitter.*

Aberdeen on the transmissions from Leafield, where the distances were respectively 184 km and 590 km and the bearings differed from each other by less than 10°.

The results of the measurements enumerated in this part of the paper indicate the existence of certain phenomena peculiar to short-distance transmission. For example, it is difficult to understand why there should be an absence of sunset variation, unless it be that the stratum of the atmosphere responsible for this high-angle deviation is different from that responsible for reflection of the waves which cause the sunset cycle observed at distances of the order of 400 to 1 000 km.

It is well known that about 100 km up the atmosphere changes rapidly, and consequently waves going above

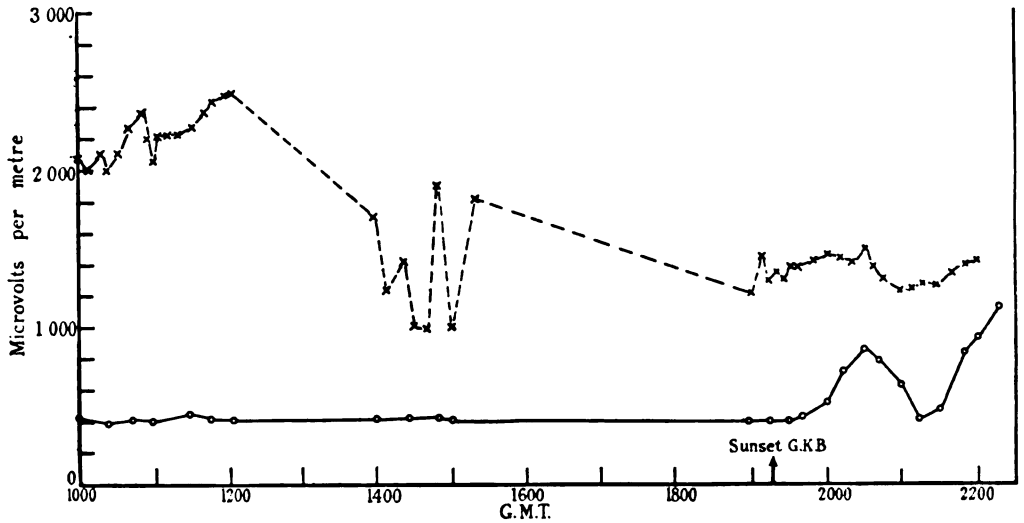

FIG. 6.—GKB (λ = 6 950 m).

- - - Manchester (distance 246 km, bearing 149). —— Aberdeen (distance 632 km, bearing 169).

In order to illustrate this effect fully, typical graphs of intensity variation made simultaneously during the day-time at Manchester (short distance) and Aberdeen (medium distance) on the transmissions from Northolt are plotted in Fig. 6. The true bearings of the transmitter from the receivers indicate that the different results can hardly be accounted for by a difference in direction of the path of the signal. The cause clearly seems to be that the part of the atmosphere from which the space wave is returned to Aberdeen has much more constant properties than is the case with the critical height for Manchester. The sunset intensity cycle is clearly shown on the Aberdeen graph.

Similar effects were obtained at Manchester and

this height will be influenced in a somewhat different manner from those which are reflected at the accepted height of the ionized layer in the day-time. At these greater heights the pressure is lower, and consequently the mean free path of the ion will be greater than at the lower altitude of 75 km usually assumed to be the height of the ionized layer in the day-time for wavelengths of 10 000 to 20 000 m. Another outstanding feature of this reception at short distances is the large apparent rotation of the plane of polarization, which gives rise to unusually large bearing variations.

This work was carried out as part of the programme of the Radio Research Board, and is published by permission of the Department of Scientific and Industrial Research.

* *Journal I.E.E.*, 1926, vol. 64, p. 584.

[The discussion on this paper will be found on page 131.]

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FIELD-STRENGTH MEASUREMENTS ON DAVENTRY 5XX.*

By R. NAISMITH, Associate Member.

(*Paper received 2nd February, and read before the* WIRELESS SECTION *1st April*, 1931.)

SUMMARY.

The paper describes a series of measurements made in different parts of Southern England on the transmissions from Daventry 5XX.

An intensive survey of Cornwall indicated a rapid attenuation of field strength on crossing Bodmin Moor. Beyond this area symmetrical contours were observed. Their existence is attributed to the large, irregularly shaped moor, rising to the hill " Brown Willy." This conclusion was further tested in Devon, where confirmation was obtained.

Experiments were made on the effects of coastal refraction, and on energy from the upper atmosphere. Negative results were obtained which proved that the effects, if any, were small.

The strength of signal received in various parts of England from Daventry 5XX has been commented upon in various publications. A fairly comprehensive series of measurements has been made in Cornwall by J. H. Reyner and published in the *Journal* (1930, vol. 68, p. 181). In this paper, two sets of contours are drawn, and it is concluded from these that the beam station at Bodmin casts a radio shadow. As the height of the beam array is only a very small fraction of a wavelength, however, it is difficult to understand how the effect suggested could be produced, without assuming a number of coincidences. The phenomenon of the radio shadow has been observed on shorter waves by Bown, Martin and Potter.† In this case a group of high buildings close to the transmitter produced the necessary absorption in a given direction. Interference effects were then clearly measured beyond this shadow wall.

The existence of other radio shadows, due to hills, masts and various other excrescences in the path of the wave, is very well known. In Reyner's paper the intensity measured in Cornwall is compared with a theoretical value which appears to have been obtained from the Hertzian formula. It must be remembered that many assumptions were implied in the application of this formula. In particular, the wave is assumed to have travelled over a flat, perfectly conducting plane. These conditions certainly do not exist in the path of the wave between Daventry and Cornwall. It would appear necessary to emphasize the importance of the curvature of the earth when dealing with distances of the order of 300 to 400 km. A simple calculation is made in the Appendix to show that if an imaginary straight line were drawn from the transmitter to the receiver, the surface of the earth would be 7 000 ft. above the centre of this line. This makes it clear that the waves travelling to Cornwall must be considerably diffracted. The formula of Macdonald† for the " Diffraction of Electric Waves Round a Perfectly Reflecting Obstacle " provides the basis for a comparison with the measured results. All the theoretical curves shown are derived from this formula.

It must be noted that part of the densely populated areas of Bath and Bristol and some quite hilly country in the south-west lie in the path of the wave. Also, open sea and thickly wooded areas intervene. These factors all introduce an attenuation effect which it would be extremely difficult to assess with any degree of accuracy.

These considerations suggested the desirability of further measurements. The apparatus used was a modification of that described by J. Hollingworth in the *Journal* (1923, vol. 61, p. 61). The modification consisted in the use of one of the resistances (used in the determination of the high-frequency resistance of the coil) as a means of introducing a known signal into the receiving coil, after the manner described by C. R. Englund.‡ The combination of these two systems of measurement ensured a greater number of accurate measurements being made than could have been done with either system alone.

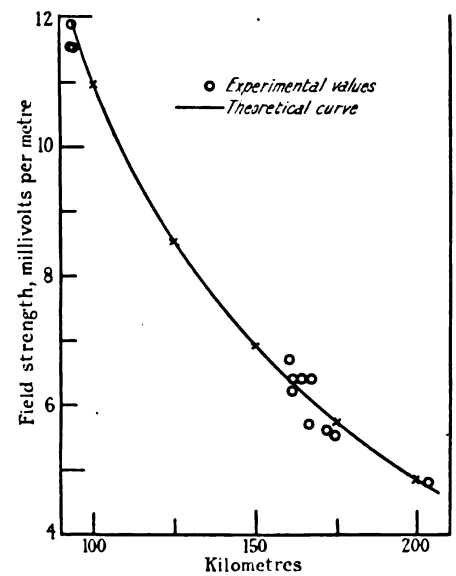

FIG. 1.—Attenuation curve.

In order to eliminate as many unknowns as possible

* Reprinted from *Journal I.E.E.*, 1931, vol. 69, p. 881.

† *Proceedings of the Institute of Radio Engineers*, 1926, vol. 14, p. 57

† *Philosophical Transactions of the Royal Society*, 1910, vol. 210, p. 113.

‡ *Proceedings of the Institute of Radio Engineers*, 1923, vol. 11, p. 26.

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from the experiment, a series of measurements were made in the east country between Ipswich and Yarmouth. This provided a path for the wave in which neither large towns nor hills higher than 300 ft. intervened. These results are plotted in Fig. 1. Observations were also taken at closer ranges but in different directions from the transmitter; these also are plotted. It can readily be shown that a calculation based on the observations at 90 km distance gives an effective height of but little more than half the actual height, unless an unusually high value of attenuation is assumed. This appears to be rather low for an aerial with such a large horizontal portion at the top. The value of 11·9 mV/m obtained at Slough, on which the effective-height calculation is based, was actually the mean value obtained on three different systems of measurement which happened to be available. The agreement between the three systems was within 5 per cent of this mean value.

The second part of the experiment was conducted in the vicinity of the Dorchester beam station, where a series of measurements were made to compare with those taken round the Bodmin beam station. The results obtained were quite consistent and nothing abnormal was measured anywhere in the locality. The average value was 4·1 mV/m, and this is about 15 per cent lower than the measurements at corresponding distances over the more ideal east country referred to above. A series of measurements were also made between Exeter and Barnstaple, and the average result for the district is plotted in Fig. 4.

Two tests were made on arrival in Cornwall to determine to what extent reflected energy might be responsible for abnormal signal strengths. By the co-operation of the British Broadcasting Corporation the frequency of Daventry was changed in nine successive steps of one kilocycle each. A measurement of intensity was made on each frequency. The calculation in the Appendix shows that this change would be sufficient to produce one complete interference cycle between a wave diffracted over the surface of the ground and one reflected from the ionized layer. As a result of this test it was clear that, at least during the day-time in the summer, sufficient energy was not received from the ionized layer to account for any abnormal signal intensities received in Cornwall. The test was carried out near Bodmin, about 305 km from Daventry.

The second test was made over the sunset period but no evidence of any change of intensity was observed, thus confirming the first test. It was possible during this experiment to rotate the receiving coil and note

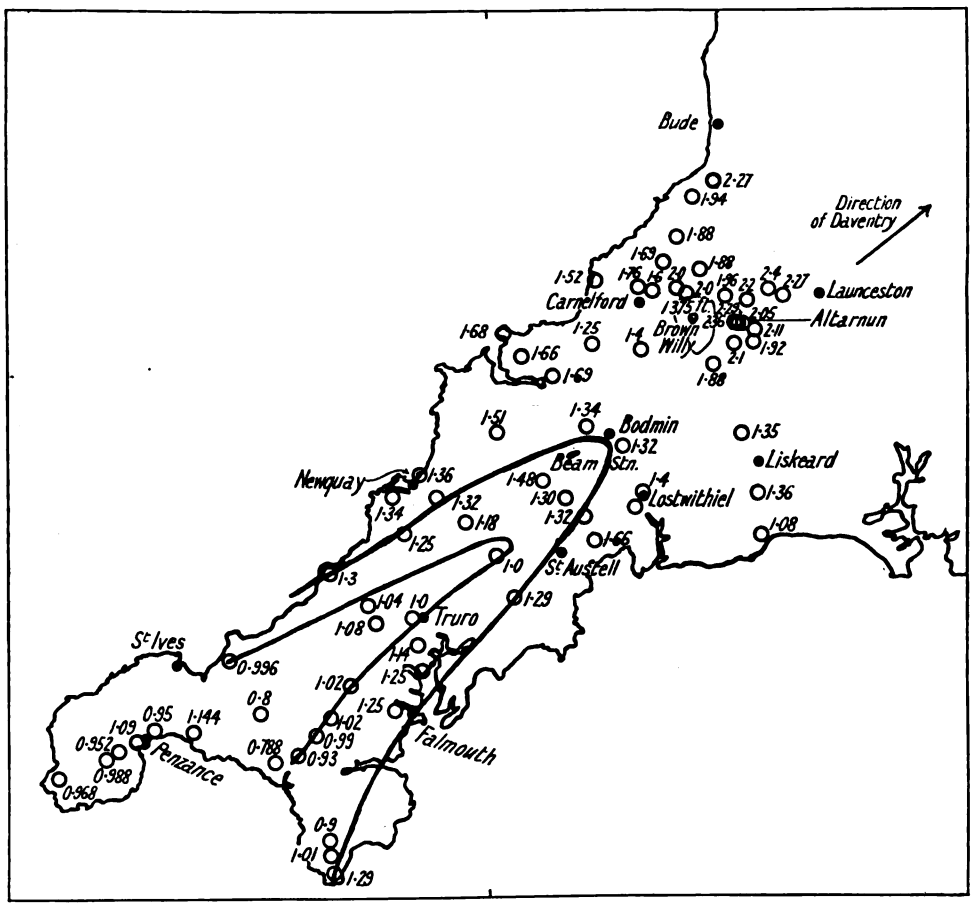

FIG. 2.—Map of Cornwall, showing contours. Measurements in millivolts per metre.

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any change either in bearing or in flatness of minimum. This check only served to confirm the conclusion arrived at after the first test.

Survey of Signal Intensity in Cornwall.

Opportunity was taken in this locality to make a number of measurements in the immediate vicinity of the Bodmin beam station, but these measurements still gave no indication of anything really abnormal. A survey of the country was therefore begun by crossing from coast to coast while maintaining a constant distance from the transmitter. Five sets of measurements were taken on this plan, each set being taken at a different distance from the transmitter. It was soon noticed that the intensity in the centre of the country was between 70 and 80 per cent of the intensity at the corresponding point on the coast. The consistency with which this ratio was maintained is the only excuse for drawing the rather striking contours. The results obtained are all set out on the map in Fig. 2. The position of the point indicates as nearly as possible the site at which the measurement was taken. It will be observed that the contours extend several kilometres beyond the beam station in the direction of the transmitter. The high value of 1·48 mV/m obtained S.W. of the beam station was checked. Compared with other values taken in the neighbourhood, it appeared to be slightly abnormal. The observing site was on the edge of the Tresayes Downs overlooking the beam station and providing an optical path extending beyond Bodmin Town. The position of this observation is considered to be of importance in view of the later results.

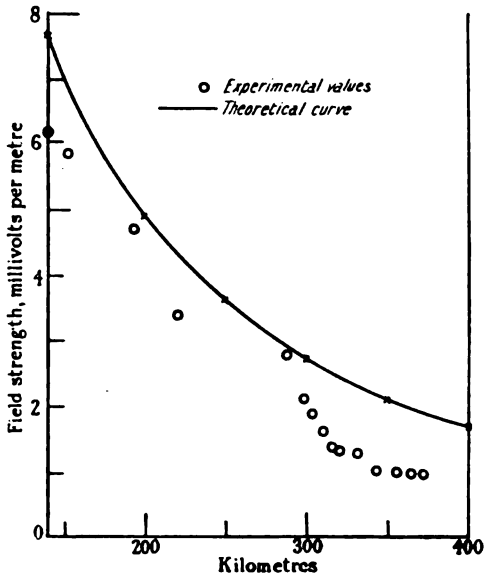

Fig. 3.—Attenuation curve.

Observations were continued over Bodmin Moor in the direction of Launceston, but it was soon apparent that a very rapid increase in attenuation had taken place near Bolventor. The actual measurements taken along the great circle which extends through Launceston and Daventry are plotted as an attenuation curve in Fig. 3. A theoretical curve which allows only for curvature of the earth is drawn. No coefficient of attenuation has been inserted. There appeared to be nothing to account for this sudden drop in signal strength, except a hill about 1 400 ft. high (Brown Willy) slightly to the North. The country over which the measurements were made was open moorland, so that it was possible to find ideal sites for the observations. A complete circuit was made of this hill and it was then evident that the previous consistency had quite disappeared and the measurements would fit neither into contours nor into attenuation curves. It was noticed that on two sites from which long optical paths extended in the direction of the transmitter, the signal strength was high. In one particular

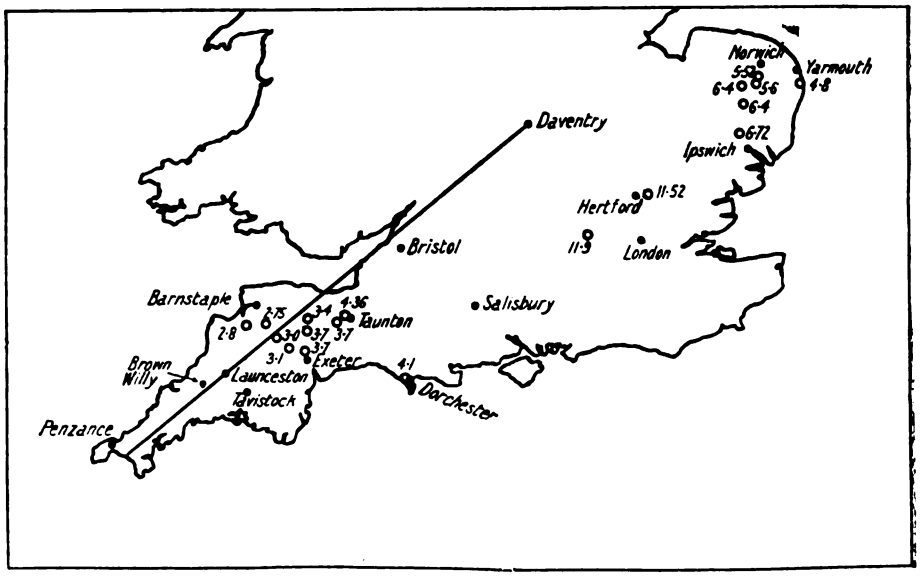

Fig. 4.—Measurements in millivolts per metre.

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case near Altarnun an increase of signal intensity of 15 per cent was observed on travelling away from the transmitter a distance of less than 1 km. A check observation taken half-way between these two sites gave an intermediate intensity value as shown on the map. Measurements on the opposite side of the hill from the transmitter were abnormally low. Further observations in this district suggested that this comparatively small hill did definitely affect the signal intensities.

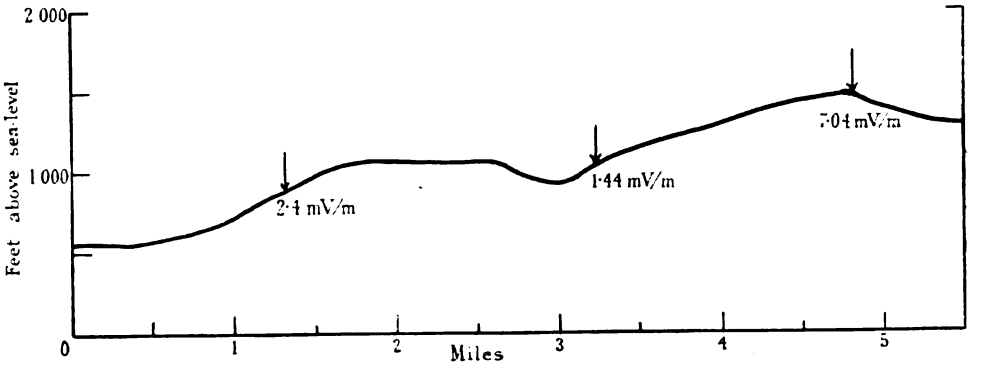

FIG. 5.—Elevation of sites in Tavistock region.

MEASUREMENTS IN DEVON.

In order to check this effect under more ideal conditions, the survey was carried on into Devon. The region around Tavistock and Two Bridges was chosen as a convenient district in which to study the effect of the Tors. One mile east of Tavistock, on the side of the hill, an intensity of 2·4 mV/m was recorded. One mile further up the hill, travelling towards the transmitter, the intensity had dropped to only 1·44 mV/m. Just over the crest of the hill, at the junction of the Moreton Hampstead and Princetown Roads, the signal strength had risen to 7·04 mV/m. About a mile west of Two Bridges the intensity was still 6·768 mV/m. At Postbridge the intensity had dropped again to only 2·8 mV/m. The highest peak in this neighbourhood is about 2 000 ft. The results on Tresayes Downs, Brown Willy

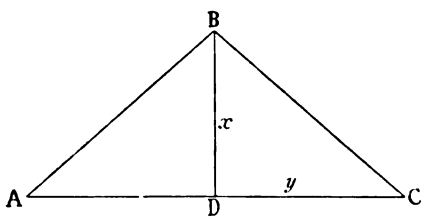

FIG. 6.

and Dartmoor all showed that when the path of the wave between the receiver and the horizon was increased the intensity of the signal was also increased. It should be noted that mere elevation of the receiver does not account for these results. For example, in the Tavistock region the intensity actually dropped by 40 per cent on climbing several hundred feet up a hill in the direction of the transmitter. This effect has already been mentioned by R. O. Cherry,* but no actual measurements are recorded. Further evidence on this subject was not sought, as the only object of extending this Cornwall tour into Devon was to verify the possibility of a hill of only 1 500 ft. affecting the field contours beyond it of 1 500-metre waves.

* *Proceedings of the Physical Society of London*, 1929-30, vol. 42, p. 192.

GEOGRAPHICAL CONSIDERATIONS.

The position of the country relative to the transmitter is worth commenting upon (see Fig. 4). A great circle can be drawn through the centre of the country and through the transmitter. The two coastlines run parallel to this great circle. The question of coastal refraction was considered and certain measurements were taken right on the sands, but it was not possible to attribute any definite effect to this cause. It is possible to state that if any coastal effect is present it is small compared with the effects enumerated in the paper.

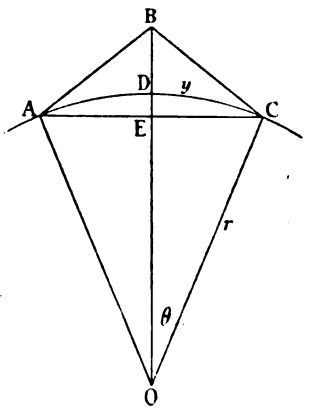

FIG. 7.

CONCLUSION.

It is concluded that the cause of the rapid attenuation west of Launceston, and the peculiarly shaped contours of signal strength plotted in Cornwall, is the large irregularly shaped moor rising to the hill Brown Willy.

The author wishes to acknowledge the assistance of Dr. Burnett in the derivation of the theoretical curves shown in Figs. 1 and 2.

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APPENDIX.

(1) *Problem of phase relation of two waves arriving at C, one travelling route ADC, the other ABC (see Fig. 6). Values taken:*

$$x = 80 \text{ km}, y = 170 \text{ km}, \lambda = 1{\cdot}554 \text{ km}$$

Waves in phase when ABC–ADC $= \frac{1}{2}n\lambda$, where n is even.

In the above case

$$n = \frac{4}{\lambda}\left[\sqrt{(x^2 + y^2)} - y\right]$$
$$= 46$$

therefore the waves are in phase.

The waves will again be in phase when $n = 48$, i.e. when ABC–ADC $= \frac{1}{2}n\lambda$,

or
$$\lambda_1 = \frac{4}{48}\left[\sqrt{(x^2 + y^2)} - y\right]$$
$$= 1{\cdot}49 \text{ km}$$

Therefore the wavelength change necessary to produce one complete interference cycle is from 1·554 km to 1·49 km (equivalent to 8 kilocycles per sec. change in frequency).

(2) *Effect of curvature of earth over ground distances of about 340 km (see Fig. 7).*

$$y = r\theta \quad \therefore \theta = \frac{y}{r} = 1^\circ 32'$$
$$\text{DE} = r - \text{OE} = r(1 - \cos\theta)$$
$$= 2{\cdot}284 \text{ km}$$

DISCUSSION BEFORE THE WIRELESS SECTION, 1ST APRIL, 1931.

Dr. R. L. Smith-Rose: With regard to the second paper, it is interesting to compare the results obtained by the author in Cornwall when making field-strength measurements on Daventry 5XX with those given in the recent paper* by J. H. Reyner. A comparison between Fig. 3 of Reyner's paper and Fig. 2 of the author's leads one to the conclusion that either the radiation from Daventry has altered considerably since the earlier measurements were made, or—what is more likely—one of the two investigators has made serious errors in his measurements. On page 128 the author hints that the accuracy of his measurements is of the order of 5 per cent, and as Reyner at the end of his paper makes a 4 per cent correction to the experimental values, I presume he also regards his results as correct to, say, 5 per cent. Reyner's values for the field strength in the neighbourhood of Bodmin are more than twice the author's figures, and the difference in shape of the contours shown in the two papers is considerable. The author draws the conclusion that the effects which he observed in Cornwall are due to the hill known as " Brown Willy," which is 1 400 ft. (0·25λ) high. I find it difficult to conceive that a hill of this height could cast over the rest of Cornwall a shadow persisting for some 50 miles. I should have thought that the dimensions of this hill would have been insignificant compared with the distances over which the measurements were taken in the south-west of Cornwall. Reyner, on the other hand, attributed his peculiar results to the effect of the Bodmin beam station. Although the height of a beam-station aerial is small (about 0·1λ), the span of a beam system is considerable. The total span of the two banks of the Bodmin aerial system is about 6 000 ft. (just over a wavelength), and while I should not like to rule out the possibility that this aerial may be producing effects upon the field intensity in other parts of Cornwall, I feel doubtful about either explanation. One might suppose that the subsoil of that part of Dartmoor over which the transmission has to pass before it reaches Bodmin would produce some effect, but on the other hand we have hitherto always considered that so long as the subsoil was covered by a depth of soil sufficient for vegetation to exist, the relatively high conductivity of this surface soil would screen the subsoil and prevent it from having any action upon the waves as they passed over the surface. It is possible that the broken nature of the ground in that part of Devon and Cornwall may be the cause of all the effects observed. On page 130 the author refers to measurements made at a number of places which are not indicated on the map, and it would perhaps be advisable to indicate their positions for the benefit of those who are not familiar with the district in question.

I disagree with the opening sentence of the first paper, which states that there has hitherto been no evidence of the influence of downcoming waves from the upper atmosphere in the daytime on long wavelengths at distances up to 300 km. In the first place, Hollingworth's measurements which were made in this country (in daylight) and published in 1926* show a curve with a dip and a maximum in it, instead of a straightforward attenuation curve. The only explanation which was forthcoming for this curve was that it was due to interference between the ground wave and a downcoming wave reflected from the Heaviside layer. Additional evidence of the same phenomenon is supplied by some of my own work on direction-finding.† Fig. A shows daily bearings on the transmissions from Leafield (λ = 12 400 m) taken at Slough at 2 o'clock every afternoon over a period of 15 months. During the winter months appreciable variations in the bearings due to the effect of the reflected waves from the Heaviside layer in the daytime were obtained. The dotted horizontal line through the middle of the results represents the true bearing, and the observed bearings are almost equally displaced on either side of this, except during the summer when there appears to be a deviation of the bearing in a clockwise direction. Thus my experimental results in the summer time may be taken as a partial confirmation of the author's observation that on Rugby transmissions (λ = 18 000 m) the bearing was always rotated in a clockwise direction. This conclusion might, however, be vitiated by the possibility of a permanent error in bearing of a few degrees due to local conditions.

* *Journal I.E.E.*, 1930, vol. 68, p. 181; also *Proceedings of the Wireless Section*, 1930, vol. 5, p. 37.

* *Journal I.E.E.*, 1926, vol. 64, p. 584; also *Proceedings of the Wireless Section*, 926, vol. 1, p. 57.

† Department of Scientific and Industrial Research (Radio Research), Special Report No. 5, 1927.

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The author's paper demonstrates that, even though a transmitting station may be radiating equally in all directions, the transmission of energy in the north-south direction is greater than that in the east-west direction at short distances. I think it is dangerous to conclude, however, that because more energy is returned in the north-south direction at short distances, less is available to be transmitted at longer distances. It does not follow that because the energy has not come down at a short distance it will come down a long way away; it may be entirely absorbed. With regard to Figs. 3 and 4, I question the utility of inserting the Austin-Cohen values. I should like to express my high opinion of the results obtained, because I am familiar with the great difficulties there are in making consistent measurements, not only at a fixed station but also when touring the country with apparatus which must be transportable.

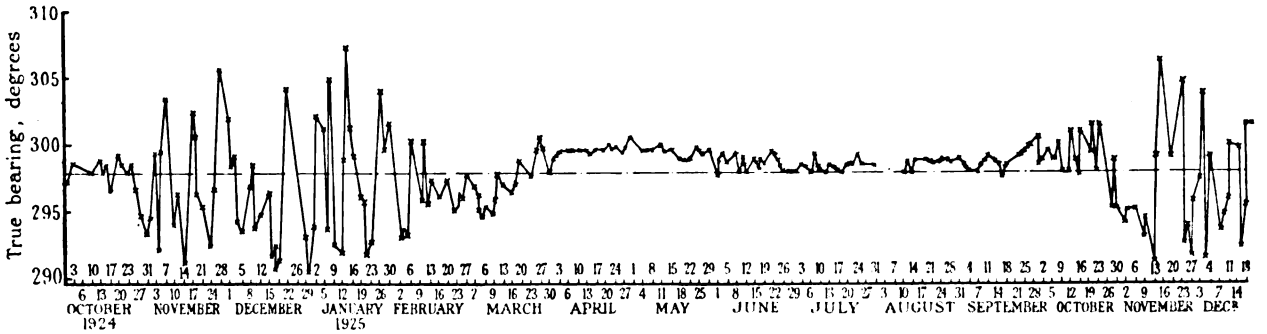

FIG. A.—Graph showing daily observations of bearings taken at Slough on the U.R.S.I. transmissions from Leafield, October 1924 to December 1925.

$\lambda = 12{\cdot}4$ km, transmissions at 14 00 G.M.T.

Mr. J. H. Reyner: With regard to Dr. Smith-Rose's remark concerning the 4 per cent accuracy mentioned in my paper,* this figure was an estimated attenuation obtained from a theoretical formula, which fact I allowed for when I was determining the absolute signal strength of Daventry. My measurements gave a signal strength of 13 millivolts per metre, which is in satisfactory agreement with the author's figures. While I am a little disappointed at the somewhat serious discrepancy between the author's and my own results in Cornwall, it should be remembered that my measurements were made entirely on my own initiative as a summer-holiday task, that I was only able to spend about three weeks on the work, and that I used very simple apparatus. In the circumstances I am relieved to find that the author did definitely notice a drop in signal strength in the interior of the county as he crossed from coast to coast. My apparatus consisted of a small portable aerial set up on each site. The signal strength from Daventry was allowed to produce a given deflection in the detector circuit; I then changed over to a local oscillator and adjusted this to give the same deflection. To do this I had to use a dummy aerial, because I could not induce the signal in the main aerial while I was still receiving a signal from Daventry. So far as I can see, this is the point where the discrepancy was introduced—although I endeavoured to anticipate and avoid any such error.

* *See* footnote on page 131, col. 1.

As to the question of whose measurements are right, I have no hesitation in accepting the author's, because his facilities are undoubtedly better than mine were. There is one question I should like to raise, however. If the unusual effect observed is caused by a radio shadow of some natural phenomenon like "Brown Willy," ought not the signal-strength contours to close again rather in the manner which I, in my ignorance, showed them to do? My original paper suggested that if there were a shadow it should close up, and a band of high pressure should be formed again before the signal strength fell off. I looked for this and found what I thought was evidence of its existence, but my measurements were taken in a short time and did not, I think, number more than 40, whereas the author has taken very many more. Purely from a theoretical point of view, I should like to know whether this closing of the contour and subsequent slight rise in signal strength ought not to be observed.

Dr. J. Hollingworth: With regard to the first paper, if we consider long waves to be merely channels of communication we probably know everything about them that it is necessary to know. Regarding them as electrical phenomena, however, there are a great number of details which are still crying out for investigation. It may seem rather peculiar that there should be any reason for a sharp distinction between the kinds of phenomena observed above and below 300 km. This distinction, however, is a very important one, and it is emphasized by the following short report dealing with some of the subsidiary phenomena mentioned in the paper by myself to which Dr. Smith-Rose has referred. In connection with a journey north on which I was measuring the signal strength from Rugby and various other stations, the report states: "The results on Rugby yielded a new factor of importance. Up to a distance of 340 km from the transmitter the results, though somewhat irregular, showed nothing very unusual; but at the next reading, only 55 km further north, the intensity dropped from 8 000 to 2 800 microvolts per metre, round about which value it remained for the rest of the journey, and fitted in very closely with the routine figures being obtained at Aberdeen. In order to see if this was merely a temporary phenomenon, I returned about a fortnight later by an entirely different route. The same type of effect was observed, signals rising from an intensity of

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5 300 microvolts per metre at 345 km to 10 000 at 305 km." This is very strong evidence that, due to some cause of which we do not know, at 300 km there is a complete change in the phenomenon, and there is reason to believe that it is fundamentally a change of quality rather than of quantity. It is difficult at present to see how to explain this change, but we should guard against the danger of trying to force the facts to suit existing theories. It is more important to recognize that this definite experimental effect occurs; we shall probably have to turn to the mathematical theory for an explanation, and this presents a difficult problem. At one time I was working on long waves, because I thought they were more particularly adaptable to mathematical estimate, but I am now told that the mathematical theory of short waves is the simpler. It is also rather striking that whereas these short- and long-distance phenomena seem to be somewhat alike, the middle-distance ones are entirely different.

Turning to the second paper, the field work involved in these investigations must have presented great difficulties. It was at one time considered that in long-wave measurements at any rate one could obtain reliable readings provided that there were no rivers, railways, or telegraph and telephone lines in the immediate neighbourhood, but these results show that there is some other cause of inaccuracy either above or below the ground.

Captain P. P. Eckersley: Measurements of effective height of aerials taken by the B.B.C. showed certain curious ambiguities. At a point about 10 km from Daventry, measurements gave an effective height of 100 m, indicating a radiation resistance greater than the resistance of the aerial. At 30 km from Daventry the effective height had apparently been reduced to about 80 m. It is interesting to note on page 128 of the second paper that the effective height (not actually mentioned) was half the actual height. The figure obtained would, I suppose, be about 60 m, which is very different from our readings near Daventry. We have frequently observed that the effective height measured close to an aerial appears greater than it should be according to theory. Incidentally, the effective-height reading mentioned in the paper will have to be multiplied by 0·9 (assuming the conductivity to be 10^{-13} C.G.S. units), on account of the slight attenuation introduced between the points of initial radiation and field measurement. I should like to know on what theory the curve shown in Fig. 3 of the second paper is based. Is it the Sommerfeld, Austin-Cohen, or the curvature theory? It is interesting to find that, on the average, the author's results in Cornwall closely approximate to figures obtained by methods outlined in my paper on " The Design of Transmitting Aerials for Broadcasting Stations."* The Sommerfeld theory, as first adapted by Mr. Barfield, is basic to the methods outlined in my paper. In making a study of the effect of hills upon field contours I found that the field on the height of land was always about 60 to 70 per cent greater than that measured on any open space in the neighbourhood. Where the land rose precipitously, the field readings showed a 200 per cent increase above those taken on the plain below. This effect was particularly noticed at Sutton Bank, Yorkshire, a large plateau ending in a precipice overlooking the plain which stretches away in the direction of Daventry. I also found that the field in a valley running in the direction in which the waves were travelling was much less than that in a valley running at right angles to the direction of wave travel. I furthermore noticed that hills caused considerable shadows, but these shadows were local and disappeared some distance behind the obstacle. As a result of a tour throughout the length of the Pennine Range I came to the conclusion that there is an increase of field on the height of land, but that " broken " country is responsible for a greater attenuation than flat ground of the agricultural type. It is noteworthy that whereas the detailed contours extrapolated by two independent observers (J. H. Reyner and the author) are noticeably different, a " general " contour arrived at on the basis of my methods for calculating the service area of broadcasting stations is confirmed by the readings taken by both investigators. It is extremely difficult to express the intensity of the resultant field except in terms of generalities; nevertheless all the work which has been done on the subject since the publication of Mr. Barfield's original contribution is useful since it expands our knowledge and helps us to understand the reasons for the variations of field from a general mean value.

Mr. H. L. Kirke: In connection with the second paper, the B.B.C. recently carried out some work over similar country in Somerset on shorter waves (300 m), for which the effect is considerably more pronounced. When prospecting for a site for the Western Regional station we found that on the top of the Brendon Hills the attenuation for the first 7 miles was far greater than we had ever experienced before. On a wavelength of 300 m over ordinary country one expects to obtain an inverse-distance law, for the first few kilometres at any rate. Possibly at 7 miles an additional attenuation of 10 per cent is to be expected, but in this particular case the wave was attenuated to one-seventh of the predicted value. On referring to a geological map we found that the point where the rapidly attenuating curve gave place to a curve of reasonable form was where the subsoil changed from Devon red sandstone to a soil which was apparently of greater conductivity. As we advanced in the direction of Plymouth we encountered the Dartmoor forest with its subsoil of granite and the signals disappeared completely. It seems to me, therefore, that the effect noticed by the author in Cornwall may have been due to the subsoil, and the reason why the intensity was greater nearer the coast is that the signals come in from the sea. " Brown Willy " probably produced only a local effect. Fig. B shows the attenuation plotted in terms of field strength × distance, so that, apart from additional attenuation due to losses in the ground, the curve should be a straight line. The actual curve changes its general direction at about 8 km, where it becomes fairly straight. At this point (see Fig. C) the sandstone ends, and the signal strength disappears completely where the granite starts. Fig. C is taken from the Geological Survey map. The shadow effect of hills seems to be very similar to that caused by a heavily attenuating piece of ground, or a town. We

* *Journal I.E.E.*, 1929, vol. 67, p. 507; also *Proceedings of the Wireless Section*, 1929, vol. 4, p. 51.

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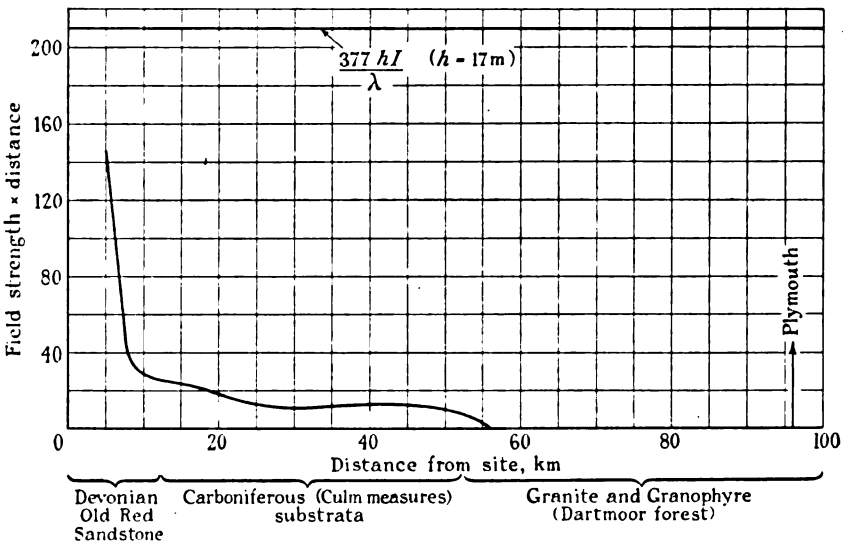

FIG. B.—Attenuation curve, Brendon Hills—Plymouth. $\lambda = 261$ m.

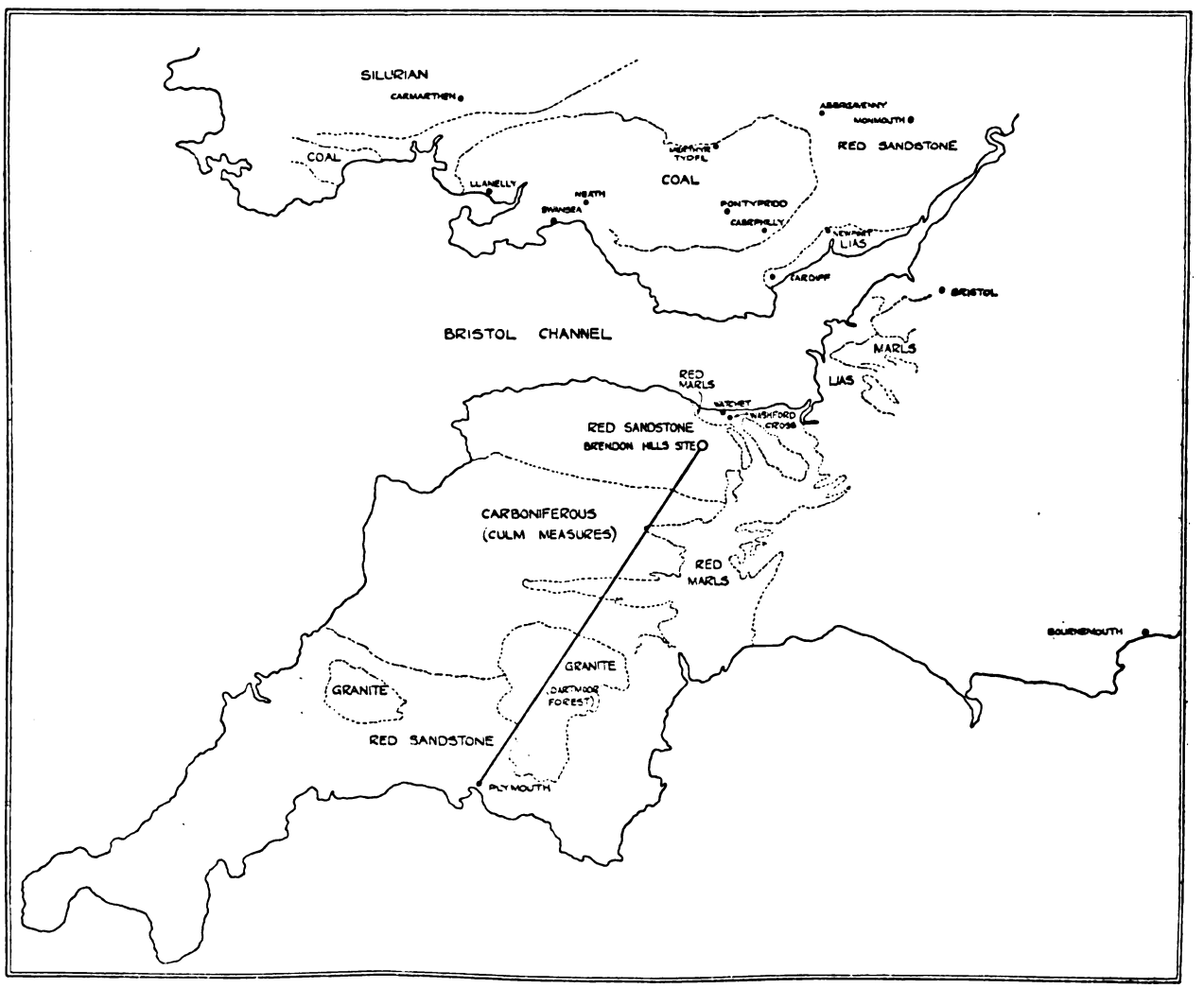

FIG. C.—Western Regional transmitter. Diagram showing geological strata encountered by direct line from Brendon Hills site to Plymouth.

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found that after waves had passed over such a region there was always a tendency to fill up. This phenomenon has been observed by Van der Pol and others, and is to be expected on theoretical grounds.

Mr. A. J. Gill: Dealing with the second paper, what surprises me about these field-strength measurements in Cornwall is not that the results of the two observers are so different, but that they agree so well. J. H. Reyner's results apparently represent one or two readings taken at each spot, and I understand that the author also only took one reading at each point. Fig. 3 of the first paper shows that whereas on one day the field strength of a long-wave station is about 1 000 microvolts per metre at a certain point, a little later the intensity at the same point has risen to 3 000 microvolts per metre. This represents an increase of 300 per cent obtained by the same observer at the same spot, and we should therefore not be surprised at a difference of 300 per cent between the results of two independent observers obtained at different times and in different places on a much shorter wave. It seems doubtful whether the assumption is justified that there was no downcoming wave, as only two tests were made at the commencement of the work. When taking field-strength measurements on the road I have noticed that a movement of only 100 yards will cause a large difference—sometimes of the order of 50 per cent—in the reading obtained. Field-strength measurements taken as one moves away from a transmitting station do not give rise to a smooth curve. A series of peaks and valleys is obtained, and a curve has to be drawn through these in order to give some idea of the attenuation. I therefore think that the readings obtained by the author cannot be average values.

Lieut.-Col. H. P. T. Lefroy: As regards the signal strength from a given transmitting station observed at any given point, previous speakers have commented upon the effects of absorption in, and interference by reflection from, the Heaviside layer, and also of screening on, and dissipation in, the earth's surface. I suggest that it may prove necessary to take into consideration the effect of minerals underlying the point of observation, and perhaps even the whole path of the radiation measured. Certain phenomena seem to indicate that there is some hitherto unobserved radiation, or field of force, associated with all minerals and characteristic of each. In some cases this suggested effect may reduce (by some form of interference) the signal strength at points situated over mineral deposits, even when the latter are at a considerable depth. Chapman and Franklin* have recently developed a method of geophysical prospecting, based upon the observed variation of signal strength in the area under investigation, which may in time provide data in confirmation of the hypothesis suggested.

Mr. R. Naismith (*in reply*): The disagreement expressed by Dr. Smith-Rose with my opening sentence on page 121 of the first paper is not confirmed by his quotation. I have defined the short distances in that sentence as distances under 300 km. The extreme limits of distance in the curve quoted lie between 300 and 1 000 km; therefore the quotation is not applicable. I would, however, draw his attention to the words "in the summer"* which, though inadvertently omitted from the advance copies of the paper, are obviously implied throughout the paper. For instance, during winter even medium-distance transmission is subject to large daily variations and the comparison in the paper would then be impossible. I agree that it does not follow that because the energy has not come down at a short distance, it *will* come down a long way away, but my statement in the penultimate paragraph of page 124, col. 1, is that if a large amount of the energy from a transmitting station *is* returned close to that transmitting station, there will be less to reach the greater distances. These two statements clearly do not coincide. The values marked A and B in Figs. 3 and 4 are simply calculated values based on the formula $E_g = 120\,\pi Ih/(d\lambda)$.

In discussing the second paper Dr. Smith-Rose finds it difficult to conceive that a hill 1 400 ft. high could cast a shadow persisting for 50 miles. The following extract from a paper by Ollendorff† on the subject shows that the shadow zone could still be observed at 4 times this distance. A model of a range of hills is represented by a semicircular cylinder, Fig. D. The waves are propagated perpendicularly to the direction of the range so that the range of hills always lies in a wave surface. If Z is a diffraction zone within which the diffracted wave shall be at least 10 per cent of the primary wave and ϵ is the dielectric constant of the obstacle, h the height of the obstacle and λ the wavelength considered, the following relation holds:—

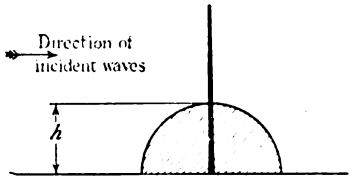

FIG. D.

$$\frac{Z}{\lambda} = (20\pi^2)^2\left(\frac{\epsilon - 1}{\epsilon + 1}\right)^2\left(\frac{h}{\lambda}\right)^4$$

Taking $\epsilon = 45$‡ and putting $h = 0{\cdot}27\,\lambda$ we get a diffraction zone of $220\,\lambda$ or 212 miles. Regarding the accuracy of measurement, I know of no standard of reference in this country, and owing to this regrettable circumstance refrain from further comment. I cannot agree that broken ground or variable subsoils could produce the symmetrical contours of Fig. 2.

In reply to Mr. Reyner, it is true that a complete exploration of the shadow should theoretically reveal certain closed contours, but as this effect has already been discussed at some length in the paper by Bown, Martin and Potter referred to in the text, I did not feel justified in undertaking the extra work which would be involved to complete the shadow picture.

I wish to thank Dr. Hollingworth for his valuable contribution to the subject of long-wave phenomena over short distances.

* "Applied Geophysics" (Board of Education), H.M. Stationery Office, 1931, p. 47.

* Since added for the *Journal*.

† *Elektrische Nachrichten-Technik*, 1931, vol. 8, p. 147.

‡ J. A. RATCLIFFE and F. W. G. WHITE: *Philosophical Magazine*, 1930, vol. 10, p. 667.

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Captain Eckersley raises the question of the theoretical intensity. The curve shown in Fig. 3 is obtained from the Macdonald formula for the diffraction of electric waves round a perfectly reflecting obstacle. It is interesting to learn that his measurements at Sutton Bank, Yorkshire, provide such direct confirmation of my results in Devon. Judging by the position of many receiving stations, I should say this effect is not fully recognized.

The measurements quoted by Mr. Kirke were made on 300 metres and cannot therefore be directly compared. Dr. Smith-Rose and Mr. Barfield have shown* that so long as the subsoil is screened by a layer of vegetation it exerts no influence on the waves passing over it. In the discussion Dr. Smith-Rose rather suggests that this hypothesis may need some revision. The measurements in Cornwall were made on widely differing subsoils such as serpentine, magnetite, slate, sand, granite and tin. Therefore the entire absence of any of the sudden changes in the attenuation coefficient noticed by Mr. Kirke, and the smooth nature of the contours passing over this exceedingly varied geological structure, lend no support to the suggestion that the subsoil affects the contours of signal strength.

The variation of 300 per cent quoted by Mr. Gill was a very definite case of change of ionization in the reflecting layer. The assumption of no downcoming wave in Cornwall was not only dependent on the results of the two tests but was further substantiated by the total number of observations, since the resultant of the reflected wave and the wave diffracted over the surface of the earth would give rise to a stationary interference pattern on the ground.

The many causes of variation of signal strength have not yet been fully explored. Lieut.-Col. Lefroy mentions one of these which is at present being investigated by Chapman and Franklin with a view to its practical application. A complete statement of all the causes and the relative importance of each could not be made at present, but the effect described in the paper is so prominent that other effects which may be present are unnoticeable.

* *Proceedings of the Royal Society*, A, 1925, vol. 107, p. 587.

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SOME OBSERVATIONS ON THE ORFORDNESS ROTATING BEACON.*

By R. L. SMITH-ROSE, D.Sc., Ph.D., Associate Member.

[From the National Physical Laboratory.]

(*Paper first received 3rd September, and in final form 13th November, 1930.*)

SUMMARY.

The following paper gives an account of observations which have been made during the past nine months on the Orfordness beacon which was established in July 1929 with the object of ascertaining the utility of this type of wireless aid to navigation when the observations were made by the personnel of the Mercantile Marine.

The Orfordness beacon is of the rotating-loop type which operates on a wave-frequency of 288·5 kilocycles per sec. (wavelength 1040 m), and emits continuous navigation signals with special morse characters to indicate the N. and E. signal positions. Bearings are obtainable with any type of wireless receiver with the aid of a suitable stop-watch or chronograph. At the request of the Board of Trade, the masters and operators of many ships have submitted a record of the observations of bearings which they have made on the beacon, together with information as to the position and true bearing of the ship at the time the observations were made. The paper gives a summary of the results contained in the reports submitted by about 160 ships and discusses these results in some detail. It would appear from the results analysed that with a normal type of one- or two-valve receiver, as employed in modern ships, the reliable working range of the beacon is of the order of 100 miles by day or night. Within this range under favourable conditions about 80 per cent of the observed bearings are correct to within 2°. At greater ranges bearings of a slightly lower order of accuracy are obtainable in the daytime up to distances of about 250 miles with the same type of receiver, and with more sensitive receivers the working range may be anything up to 500 miles. At these larger distances, however, the reliability of the bearings is somewhat impaired during the hours of darkness by the phenomena of flat minima and wandering zeros which are characteristic of experience in direction-finding under similar conditions.

Section (3) of the paper gives details of a series of observations which have been carried out at the National Physical Laboratory. From these it appears that the accuracy of the time of rotation of the beacon is well beyond the limit of any ordinary chronograph or stop-watch which is employed in bearing observations. From systematic readings obtained in the daytime over a considerable period, it is also to be concluded that skilled observers can obtain bearings from the beacon with a maximum error of 2°, and that a high proportion of the bearings are accurate to within 1°. These results, together with those (already referred to) taken on board ship, indicate that no definite error can be ascribed to the rotating beacon installation itself. Over the range of 93 miles overland between Orfordness and Teddington, observations taken at night show that the accuracy of bearings, even with skilled observers, seriously deteriorates and errors of as much as 20° may sometimes be experienced. Since it appears that the working range of the beacon might with advantage be increased above its normal value of 100 miles oversea, it will be desirable in future to develop the beacon along lines which will provide freedom from these night errors, which at present appear to be the only limitation imposed upon the accuracy and reliability of the bearings supplied by the rotating-loop type of transmitter.

* Reprinted from *Journal I.E.E.* 1931, vol. 69, p. 523.

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- (*d*) Coastal deviation.
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References.

(1) INTRODUCTORY AND HISTORICAL.

The development of the rotating-loop beacon system of wireless direction-finding by the Air Ministry in this country has been described in a previous publication.* In order to obtain data upon the reliability of operation of such a beacon transmitter, particularly from the standpoint of its application as an aid to marine navigation, the Radio Research Board carried out an investigation over a period of two years with an experimental beacon erected by the Air Ministry at Fort Monckton, near Gosport. The results obtained in these experiments, which have been described in detail elsewhere,† were so promising that the Board of Trade decided to seek the co-operation of the Air Ministry in erecting a rotating-loop beacon of somewhat improved form, at Orfordness, Suffolk, on the east coast of England (see Fig. 1) for the purpose of enabling the Mercantile Marine to carry out a first-hand trial of the system. Accordingly the necessary plant and apparatus was installed at Orfordness in the early part of 1929 and the beacon was set in operation in June of that year.

A brief description of the beacon, its mode of operation and the method to be adopted in obtaining bearings with the aid of the signals received from it was given in the Board of Trade Notices to Mariners.‡ As in the

* See Reference (1). † *Ibid.*, (2). ‡ *Ibid.*, (3).

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former installations, the beacon consists of a loop transmitter which rotates about a vertical axis at a uniform speed of 1 revolution per minute. At a distant receiver the intensity of the signals from the beacon varies according to a cosine law with the angle between the plane of the loop and the direction of the receiver, passing through a maximum and a minimum twice each minute. When the direction of minimum signals is along the meridian through the transmitter the beacon emits a characteristic morse signal termed the North signal, on hearing which the observer at a distant receiving station starts the seconds hand of a stop-watch or chronograph. The position of the hand is observed when the signal passes through its minimum intensity, and the interval that has elapsed between the N. and minimum signals may then be used directly to obtain the bearing of the receiver from the transmitter by a simple calculation. In order to make the operation as simple as possible it is convenient to fit the watch or clock with a dial engraved in degrees and points of the compass. A description of various types of such dial has been given previously,* and watches so equipped are already on the market.

The transmissions from the Orfordness beacon are made on a wavelength of 1 040 m (frequency 288·5 kilocycles per sec.), and during alternate hours of operation waves of Type A1 (i.e. continuous waves) and Type A2 (i.e. interrupted continuous waves) are employed.

(2) Observations Made in Ships at Sea.

The previous investigations conducted by the Radio Research Board had provided information as to the reliability of the bearings obtained from such a rotating beacon when the observations were made by a skilled observer with a receiver either on land or in a ship at sea. The first object of the erection of the Orfordness beacon was to ascertain what performance was obtainable when the observations were carried out by the personnel of the various classes of ship which would ultimately use the beacon signals as an aid to marine navigation. Accordingly the Mercantile Marine Department of the Board of Trade invited the masters of ships and wireless operators to ascertain the bearing of their ships from the beacon by the aid of the received signals, especially when their ships were in known positions and when bearings were also obtainable by visual or other means. Arrangements were made for the results to be recorded on a special form and sent to the Board of Trade for analysis. The results summarized and discussed below in this section were obtained in the above manner and were submitted by some 160 ships of all descriptions between July 1929 and April 1930.

(*a*) *Results obtained at Fixed Positions at Sea.*

Before proceeding to this discussion, it will be interesting to consider the results recorded in Table 1 of the observations obtained by the Trinity House authorities at or near the positions indicated by the various buoys and light-vessels in the south-western portion of the North Sea.

In this table it is seen that 16 out of the 21 observations recorded show no error in the observed bearing and that the maximum error in the remaining 5 cases is 1 degree. This refers to the case of bearings taken in a ship at anchor in such a position that its true bearing from the beacon is accurately known: under such favourable conditions it is to be concluded that at ranges up to 45 miles the wireless bearings obtained from the beacon are accurate to within 1 degree. This conclusion is comparable with that obtained from observations on land in daylight at a range of over 90 miles as recorded in a later portion of the paper.

Table 1.

Results obtained by the Trinity House Authorities at Various Positions, 9th July to 9th August, 1929.

Position of receiver	From Orfordness		Error in wireless bearing
	Distance	True bearing	
	miles	degrees	degrees
—	4·0	326·0	0·0
Shipwash light-vessel ..	8·5	115·0	−1·0
Outer Gabbard light-vessel ..	18·1	106·0	−1·0
Galloper watch-buoy ..	25·0	145·5	0·0
Barrow Deep light-vessel ..	25·0	201·0	0·0
Black Deep watch-vessel ..	29·0	195·0	0·0
N. Goodwin light-vessel ..	45·0	180·0	0·0
Elbow gas buoy	43·0	181·0	−0·5
Near Kentish Knock light-vessel	26·4	168·0	0.0
At Kentish Knock light-vessel	26·4	171·0	0.0
—	7·0	114·0	0·0
Outer Gabbard light-vessel	18·1	106·0	−1·0
S. Outer Gabbard buoy ..	20·0	117·0	0.0
N. Galloper buoy	21·5	135·0	0·0
Galloper light-vessel	25·0	144·0	0·0
N. Goodwin light-vessel ..	45·0	180·0	0·0
Off Margate Roads	40·0	183·0	0·0
5 miles S.S.W. of Kentish Knock light-vessel ..	30·0	174·0	0·0
Longsand Head gas buoy ..	18·0	175·0	−1·0
Galloper light-vessel	25·0	144·0	0·0
Outer Gabbard light-vessel ..	19·5	107·0	0·0

(*b*) *General Results obtained at Sea.*

In considering the results obtained by various ships at sea engaged upon their normal duties, attention must be directed to various factors which may affect the accuracy of the wireless bearings recorded. In the first place it appears from the reports that the majority of the ships were not provided with a 60-seconds stop-watch or chronograph suitable for making accurate bearing observations, and that the master or operator had to resort to the use of the seconds hand of the ship's chronometer. Such an instrument is difficult to read to an accuracy of within half-a-second, and also a calculation has to be made for the time of the North signal. When it is realized that the beacon rotates at 6 degrees

* See Reference (4).

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per second, it is evident that this method is subject to a possible error of ± 3°. Secondly, in order to obtain the error in the observed wireless bearing it is necessary to know the true bearing of the ship from the beacon: sometimes this can be obtained from visual observations but in other cases resort must be had to dead reckoning, with a consequent decrease in its reliability. Any difference between the wireless and the visual or calculated bearing must be ascribed to an error in the former, although previous experience has shown that this is by no means always the case. There is further the point that even when a stop-watch is employed it requires some little practice for an observer to attain the highest possible accuracy in reading the time of occurrence of the signal minimum after the N signal.

In view of these facts a general survey of all the reports received from ships indicates at once that the rotating-beacon system is successful in providing the mariner with a new means of obtaining bearings to an accuracy which is usually quite adequate for his needs. At ranges up to 100 miles no difficulty has been experienced in obtaining bearing observations with the aid of the 1-valve or 2-valve type of receiver which is now fitted to the majority of ships. In many cases the range of reception and observation was considerably greater, a matter which will be referred to below.

Table 2 gives a summary of the records sent in by those ships which have taken more than 10 observations on the beacon in either one or more series.

Although, as will be seen from the table, the range of observation was over 900 miles in one case, the majority of the results were obtained within 100 miles of the beacon and in the daytime. Also, in most of the cases tabulated a stop-watch was employed, either of the usual type calibrated in fifths of a second or of the special type already referred to in Section (1). From this mode of presenting the results it will be noted that, while in one case an error in bearing of 12° was recorded, in the majority of instances the maximum error was 4° or less. The variability in the results is undoubtedly due to the varying skill and experience of the observers, the fluctuating conditions as regards interference in the wireless reception and the sum of the personal and watch errors. Altogether, out of over 1 500 observations taken by observers on 19 ships it is found that 80 per cent of the wireless bearings are correct within 2°. This result compares very favourably with that obtained in the earlier experiments with the Gosport beacon* when, with skilled observers located in ships in accurately known positions, it was found that 85 per cent of all observed bearings were correct to within 2°. When it is pointed out that the normal type of stop-watch used beats every fifth of a second, which represents a rotation of the beacon of 1·2°, it will be appreciated that an accuracy much better than the above is not to be expected unless more accurate timepieces are employed. The limiting accuracy of 2° is comparable with that

TABLE 2.

Summary of Bearing Observations on Orfordness Rotating Beacon obtained by Various Ships between July 1929 and April 1930.

Ship	Number of observations	Distances from beacon	Maximum error in wireless bearing	Percentage of bearings with error 2° or less	Remarks
		miles	degrees	per cent	
H.M.S. "Adventure" ..	433	13	8	69	Moored in Harwich
H.M.S. "Adventure" ..	118	3	5	67	At sea near beacon
H.M.S. "Argus"	74	27 to 39	7	65	
s.s. "Bangalore"	143	43 to 908	3·5	87	2-valve receiver
s.s. "British Duke" ..	24	19 to 49	4·5	58	Single-valve detector
s.s. "British Scout" ..	15	15 to 270	4	80	
s.s. "British Tommy" ..	112	9 to 464	8	85	2-valve receiver
s.s. "City of Palermo" ..	52	8 to 110	3	79	
s.s. "Corpath"	40	12 to 160	5·5	85	
M.V. "El Argentino" ..	39	13 to 208	2	100	
s.s. "Godetia"	14	8·5 to 49	4	71	
s.s. "Isle of Thanet" ..	108	59 to 69	6	79	
s.s. "Jabiru"	55	45 to 228	6	65	1-valve and crystal detector
s.s. "Panama Transport"	78	55 to 291	4	94	2-valve receiver
H.M.S. "Pangbourne" ..	11	41 to 57	2	100	3-valve receiver
s.s. "Perth"	14	9 to 77	2	100	
H.M.S. "Renown" ..	19	27	7	79	
H.M.S. "Rosemary" ..	79	30 to 550	12	72	5-valve receiver
s.s. "Suecia"	58	9 to 196	3	97	1-valve detector and amplifier
Train Ferry No. 1	21	13 to 78	4	95	
Total	1 507	3 to 908	12	80	

* See Reference (2).

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of the normal ship's wireless direction-finder, and is, further, considered to be adequate for most navigational purposes. Apart from the possibility of a small coastal deviation error in certain directions, there is no evidence amongst all the results reviewed that any appreciable instrumental error is associated with the rotating beacon itself, the accuracy of rotation of which is considered to be superior to that of the timepieces employed in the observations.

(*c*) *Range of the Beacon for Reliable Working.*

In the report on the Gosport experiments carried out on a wavelength of 525 m it was stated that the reliable working range of that beacon for first-class bearings was about 50 miles with a normal type of ship's 2-valve receiver, while slightly less reliable bearings were obtainable up to about 100 miles by day or night. The Orfordness beacon is evidently much more efficient in regard to its radiated power,* for the working ranges of the beacon to the same classes of receiver are about double the above values. This is partly due to the greater suitability of the wavelength of 1 040 m.

The alternate transmission of interrupted continuous waves from Orfordness was intended for the use of ships still fitted with crystal receivers, or with valve detectors without either a retroaction adjustment or a local oscillator. From various reports received, however, it appears that in these cases the range of the beacon's signals for bearing observations is less than 10 miles, and its utility is therefore doubtful. With ships employing valve sets for continuous-wave reception the range at once becomes 50 to 100 miles for a 1-valve receiver (cf. Table 2). This range for accurate bearings appears to be seriously limited in the North Sea by interference from Dutch broadcasting stations on a wavelength of about 1 070 m.

With a receiver consisting of a single detector with retroaction the working range of the beacon is at least 50 miles, while if a second amplifying valve is employed this range is increased to about 250 miles, either overland or oversea. For example, s.s. "Glen Gower" when at various fixed points in the Bristol Channel obtained good bearings on the beacon at distances of from 180 to 240 miles. With the aid of a standard type of 5-valve receiver, H.M.S. "Rosemary" was able to take a consistent and moderately accurate series of bearings at distances of from 350 to 550 miles from the beacon in an approximately northerly direction.

The reports contain several interesting records of "freak" reception of the beacon at exceptionally great distances. The most notable of these is that submitted by s.s. "Bangalore." When in a position near Algiers at a distance of 908 miles from Orfordness and using a single-valve detector only, the operator took a series of seven bearings, the mean of which was practically coincident with the true bearing. In this instance, however, the arc of zero signal was approximately 15° or 2·5 seconds on the stop-watch, and the accuracy

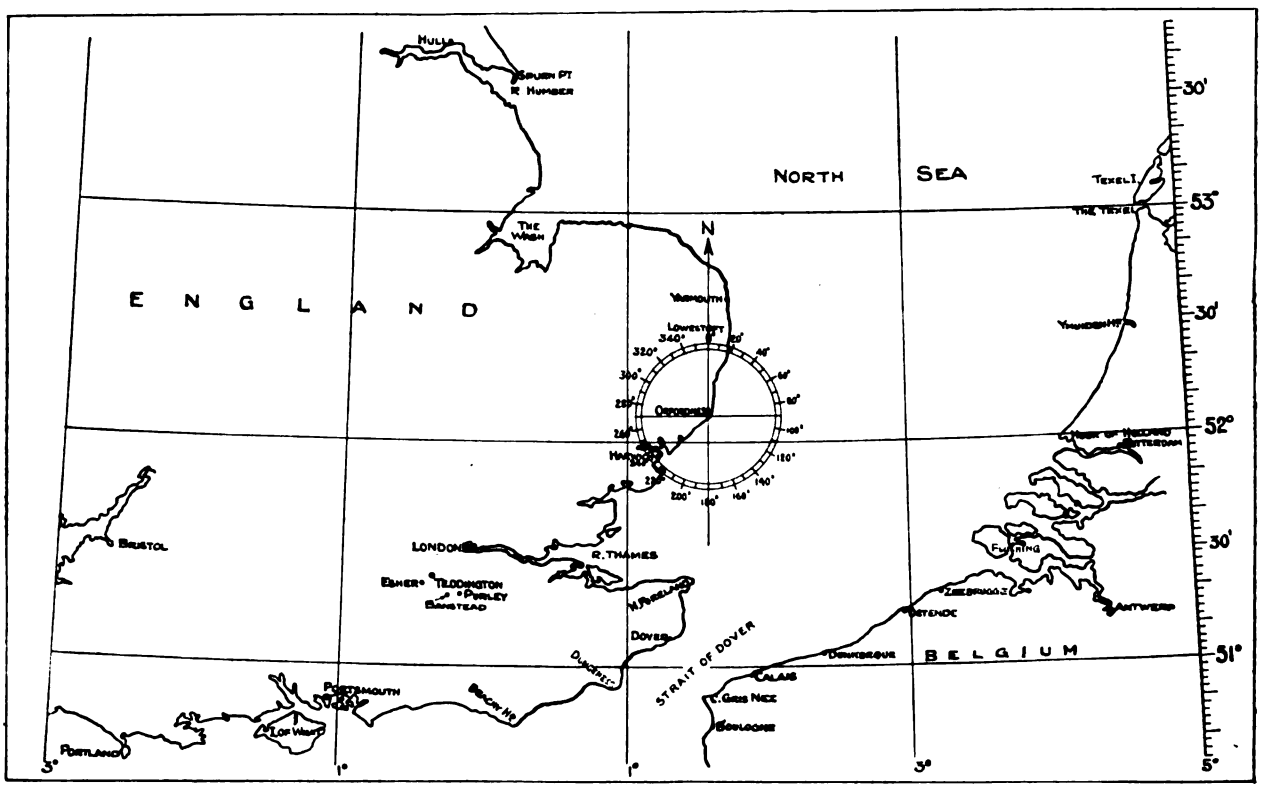

FIG. 1.

* The transmitting loop at Orfordness consists of 4 turns on a rectangle of sides 11 ft. by 10 ft., and the radio-frequency current in this loop is about 80 amperes.

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of the observed bearings must have been largely due to the favourable conditions prevailing at the time they were taken. Another instance is that of s.s. "Sasebo" which, when fishing off the West coast of Ireland, took bearings on the beacon with a 2-valve receiver at a distance of 491 miles. In a similar manner s.s. "British Tommy," using a 2-valve receiver, was able to obtain bearings at a distance of 464 miles to an accuracy of from 0·5° to 8°. The s.s. "Hebrides" has also submitted bearings accurate to within 2° taken while the ship was at anchor at distances of 428 and 446 miles from the beacon.

(*d*) *Coastal Deviation.*

When a series of experiments was carried out several years ago with a direction-finding station situated at Orford, about a mile inland from the present beacon, it was found that in the directions 225° to 240° the observed bearings were subject to a permanent error which amounted to 2·5° on a wavelength of 1 000 m.* This error was attributed to a coastal deviation effect, since in the directions in which it occurred the waves travelled practically parallel to the coast-line. It is natural, therefore, to inspect the results obtained on the beacon to ascertain whether any similar error is present in the bearings obtained from it. Definite evidence of the existence of such an error is available in the results under discussion and also among those taken on land at a distance of about 93 miles from the beacon. The results summarized in Table 3 in the next section show that in the approximate direction 240° from the beacon a permanent error of from 1° to 2° is experienced and is subject to change with comparatively small changes in the direction of transmission.

In partial confirmation of this two ships—s.s. "British Scout" and s.s. "Mundra"—have reported that when in the Thames Estuary in a direction of from 200° to 220° from the beacon, the observed bearings appeared to be subject to a permanent deviation of 2° or 3°. Further, both H.M.S. "Adventure" and Train Ferry No. 2 experienced an error of +1° and +2° respectively when they were moored at Harwich in a direction 232° from the beacon.

Somewhat similar results in a direction north of the beacon were experienced by s.s. "Bangalore." Amongst a series of bearings which are otherwise generally accurate to within 1°, this ship submitted some taken at a distance of 43 to 44 miles in directions 18° to 28° from the beacon. This approximately coincides with the direction of the coast-line to the north of the station (see Fig. 1), and the observed bearings were found to be subject to a deviation about −2·5°.

(*e*) *Night Effects.*

The majority of the results recorded in Table 2 were taken in the daytime, and the summary thus represents the order of accuracy of bearings obtainable at the various ranges indicated. At distances up to 50 or 60 miles this accuracy will be maintained during the hours of darkness, but at greater ranges there is the possibility of the received signals being accompanied at night by variable phenomena which are commonly grouped under the term "night effects." In the first place there is the possibility of fading of the received signal taking place and introducing false minima; secondly, the signal minima become blurred or even non-existent in extreme cases; while on other occasions sharp signal minima may be obtained, which, however, give bearings considerably in error by variable amounts. All these occurrences are due to interference effects between the waves received along the earth's surface and others which have travelled from the beacon through the upper atmosphere. It is known from previous experience in marine direction-finding that these effects are not usually encountered at distances of less than 70 miles,* while the bearing errors do not become serious until the distance exceeds 100 miles, the transmission being oversea in both cases. When the path of transmission is entirely overland the above distances are somewhat reduced. The earlier investigation with the Gosport station confirmed these results for rotating beacons working on a wavelength of 525 m, and, on the longer wavelength of 1 040 m now employed for Orfordness, further confirmation has been obtained in the ship reports from which the following selection is made.

An interesting example is that of s.s. "Suecia," which has submitted several series of results. Most of these observations were taken at distances of less than 60 miles and the longer distance results included in Table 2 were obtained in the daytime, with the high percentage accuracy there stated. In later records submitted by this ship the errors in bearings recorded at night at distances of about 70 miles range from 4° to 15°, although on the same occasions the errors were negligible at 45 miles from the beacon. s.s. "Sasebo," which has already been referred to as having taken approximate bearings at a distance of nearly 500 miles from the beacon, has reported that on another occasion, when off Lundy Island at a distance of over 200 miles, the signals from the beacon, while strong, gave indefinite minima and were subject to fading. In a series of results sent by s.s. "Brighton" while on the Newhaven–Dieppe service at night (distance 97 to 125 miles), complaints were made that on occasions it was difficult to obtain reliable bearings owing to the flatness of the signal minima, while at other times the results were accompanied by the remark, "Good bearings, minima clear and sharp." Among the reports from s.s. "Jabiru," which, as shown in Table 2, has usually obtained moderately accurate observations, there are some results taken at night at distances of 125 to 290 miles which show errors ranging from +14° to −30°, while at other times the operator remarked that the signal minima had practically disappeared. A somewhat similar experience is recorded by s.s. "Panama Transport" which in daylight has taken bearing observations to a high degree of accuracy at various distances up to 290 miles. On one occasion near midnight, however, when the ship was in the neighbourhood of the Casquets, about 190 miles from the beacon, the signal minima are described as varying both in quality and time of occurrence.

Some further discussion of these night effects in con

* See Reference (5).

* See Reference (5).

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nection with the rotating beacon is given in Section 3(*c*) of the paper.

(*f*) *General Conclusions on Reports from Ships.*

All the reports received to date indicate that the inauguration of the rotating-beacon service at Orfordness has been a conspicuous success in the Mercantile Marine, and it is probable that a cessation of the transmissions would cause widespread regret. On a conservative basis the reliable service range of the beacon has shown itself to be from 100–200 miles with a normal 2-valve receiver on the ship. With a more elaborate receiver the working range may be increased up to 400 or 500 miles. In almost every case the ship's receiver is used either in an oscillating condition or in conjunction with a separate valve-oscillator unit. To a crystal or 1-valve receiver without reaction the range of the beacon is found to be less than 10 miles, and within this range considerable difficulty is experienced from interference by Dutch broadcasting stations, on a wavelength of 1 070 m, owing to the comparatively unselective nature of the above types of receiver. From this point of view it is perhaps doubtful whether the transmission of interrupted continuous-wave signals in alternate hours is worth while, since it is believed that in almost every case these signals are received on the heterodyne method as with the continuous-wave signals, and the range is undoubtedly greater on the continuous-wave transmissions. It is possible that the interrupted continuous-wave transmissions may facilitate searching for the beacon signals, but after the average operator has had experience in the reception of the signals it is doubtful whether the interrupted continuous-wave transmission is further required.

The results received to date show that on the whole no definite error can be ascribed to the rotating-beacon system, the accuracy being chiefly limited by that of the watch or chronometer employed for the observations and by the skill of the observer. There is some evidence throughout the reports received that there is a permanent error of the order of 2° in the sectors 18° to 28° and 200° to 240° from the beacon. While such an error can be explained as a coastal deviation effect when the waves from the beacon are passing approximately parallel to the coast in the above sectors, it is desirable that the magnitude of the error should be fixed much more accurately by systematic observations made in the sector from various fixed positions. In any case, however, it appears that this error is limited to about 2°.

At the ranges referred to above the results show that in about 80 per cent of cases the error in observed bearings due to the timepieces and observer do not exceed 2°, except under very unfavourable conditions of interference or night effect. The latter occurrence, which is characterized by flatness of the signal minima and wandering bearings, is only of serious importance at night when the distance of transmission is in excess of about 100 miles. In such circumstances the observed bearings must for the present be regarded with some suspicion. As in the case of wireless direction-finding, the prevalence of fog or other bad-weather conditions has no effect upon the observed bearings; in the case of the rotating beacon the wireless bearings are also independent of the yawing or rolling of the ship or the accuracy of its compass.

The remarks which have accompanied the reports from many ships serve to illustrate the appreciation of the rotating beacon service by the mariner. It is considered to be desirable that all ships should be supplied with suitable watches or chronometers for use with the beacon signals. Some of the reports received also indicate that a little instruction in the adjustment of receivers and the correct method of making bearing observations would be appreciated by some of the masters and operators on ships, to whom the beacon transmissions would be of considerable service.

(3) Observations Made on Land.

In addition to the records received from ships, a certain number of observations on the Orfordness beacon have been made on land, chiefly at the National Physical Laboratory, Teddington, and a brief summary of these, with a discussion thereof, is given in this section.

(*a*) *Time of Rotation of Beacon.*

The accuracy of the bearings obtained from a rotating beacon depends upon the measurement of a time interval. It is essential, therefore, that the beacon shall rotate at a very uniform speed and, in order to avoid the necessity of timing each rotation at the receiving end, it is desirable that the time of rotation shall be 60 seconds to a high order of accuracy. Some tests were therefore made at Teddington in which the rotation of the beacon was observed continuously over periods of from 1 to 5 hours: this is made possible by the fact that, although the transmissions are interrupted at 5-minute intervals, the rotation of the beacon remains unaltered. The results of these tests show that in any daily period the mean error in the time of rotation varied from 0·001 to 0·05 sec. per minute. This is well beyond the limit of accuracy of any ordinary chronograph or stop-watch, and the time of rotation of the beacon may, therefore, be taken as 60 seconds exactly, the maximum error in observed bearing which would thereby result being less than 0·3°.

(*b*) *Day Observations of Bearings.*

During the first three months' operation of the beacon, regular observations were made at the National Physical Laboratory for about one hour daily. Further observations were made in December 1929 and April 1930 in connection with special experiments then in progress. The results of these observations showed that during any daily period the maximum departure of an individual bearing from the mean value was of the order of 1 degree. A summary of the total number of 1 444 readings made over all the periods of observations is given in Table 3, and this shows that during a period of 7 months the maximum departure of any bearing from the mean was 2°, the large majority of the bearings being within 1°, which is approximately the limiting error of the stop-watches employed. It is also seen from the table, however, that the mean observed bearing differs from the true bearing by +1°. This error was partially expected as a result of coastal deviation, which has

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already been discussed in Section 2 (*d*) above. In order to obtain a little more data upon this matter, further observations were made in the daytime at three other fixed points at about the same distance from the beacon, but in directions differing slightly from that of Teddington. The results are recorded in Table 3 and show that in neighbouring directions this coastal deviation error may change in sign but does not exceed the 2° limit which is considered to be applicable to this cause.

Neglecting this permanent deviation it is evident from the results given in Table 3 that, with skilled observers, wireless bearings from the beacon can be obtained over long periods of time to an accuracy of 2° or less, when the receiver is at a fixed point on land

TABLE 3.

Summary of Day Observations made at Various Receiving Stations.

Position of receiver	From Orfordness		No. of observations	Mean bearing	Maximum departure from mean	Error of mean
	Distance	True bearing				
	miles	degrees		degrees	degrees	degrees
Teddington	93	241·0	1 444	242·0	−2·0	+1·0
Esher	95	240·3	176	239·9	−1·1	−0·4
Banstead	93	236·1	178	234·8	+1·6	−1·3
Purley	90	235·3	57	233·3	−1·0	−2·0

TABLE 4.

Summary of Night Observations made at Various Receiving Stations.

Place	Date	Time (G.M.T.)	No. of observations	Mean observed bearing	Max. departure from day mean	Percentage of observations		
						Minima too flat for observation	Within 2° of day mean	Within 5° of day mean
				deg.	deg.	per cent	per cent	per cent
Teddington ..	23.11.29	17.30	72	236·0	−21·0	56	14	22
	19. 1.30	00.00	76	242·5	− 4·0	8	89	92
	22. 1.30	16.00	96	242·5	+ 3·5	1	93	99
	18. 3.30	00.00	74	241·2	− 5·0	8	55	91
	29. 3.30	00.00	76	241·9	+ 7·5	55	13	36
Esher ..	15. 1.30	00.00	48	240·8	+ 8·0	0	63	96
	19. 1.30	00.00	96	240·4	+ 6·0	0	48	94
	22. 1.30	16.00	96	240·5	+ 5·5	1	63	99
	31. 1.30	17.00	96	240·6	+ 7·5	0	66	96
	12. 2.30	17.10	79	239·9	+ 3·5	0	90	100
	22. 2.30	00.00	85	242·8	+15·0	0	21	65
	1. 3.30	00.00	95	239·5	−18·0	0	39	74
	8. 3.30	00.00	95	239·1	+18·0	16	12	24
	18. 3.30	00.00	94	240·3	+ 7·5	0	40	80
	29. 3.30	00.00	96	238·7	+18·0	17	17	32
	12. 4.30	00.00	79	239·3	−17·0	3	28	72
Banstead ..	15. 1.30	00.00	55	233·5	− 4·0	0	78	100
	19. 1.30	00.00	92	234·0	− 4·0	0	90	100
	22. 1.30	16.00	96	235·5	+ 3·0	0	95	100
	31. 1.30	17.00	94	234·6	− 4·0	2	88	98
	12. 2.30	17.10	91	234·2	− 4·5	0	92	100
	22. 2.30	00.00	96	237·4	+12·5	0	37	75
	1. 3.30	00.00	95	235·6	− 7·5	3	58	93
	8. 3.30	00.00	95	235·7	−11·5	21	4	34
	18. 3.30	00.00	96	234·1	− 7·0	7	59	89
	29. 3.30	00.00	93	232·4	−17·0	28	14	30
	12. 4.30	00.00	91	233·0	− 9·0	18	33	68

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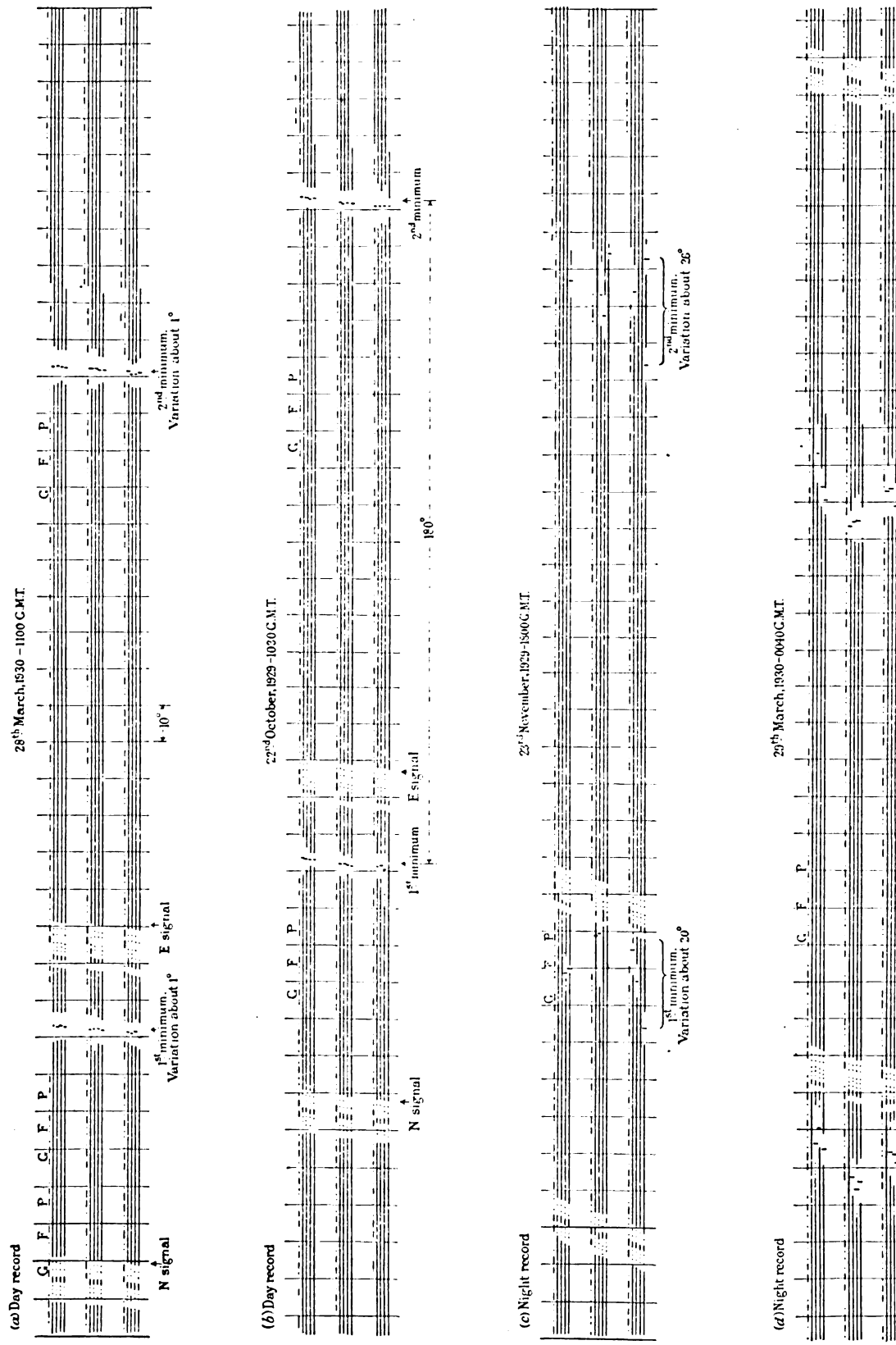

FIG. 2.—Samples of day and night records of bearings from Orfordness rotating beacon, obtained at the National Physical Laboratory, Teddington.

Notes:—The drift of the N and E signals is due to a slight difference in speed (about 0·1 per cent) between recorder drum and beacon.
Records (*a*) and (*b*) show the steadiness of the day bearings.
Records (*c*) and (*d*) show the variability of the night bearings and the occurrence of flat signal minima when the recorder pen does not operate.

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at a distance of nearly 100 miles from the beacon. Under favourable conditions a similar accuracy would probably be obtainable in a ship at twice this range.

(c) Bearings at Night.

In the previous experiments carried out with the Gosport experimental beacon to which reference has already been made,* some attention was paid to the magnitude and frequency of the variations of bearings experienced at night. It was concluded that with the rotating-loop beacon the night errors were comparable with those experienced with a loop direction-finder used under similar conditions. This work was carried out on the wavelength of 525 m on which the Gosport beacon was operated during the majority of the tests. On the working wavelength of 1 040 m for the Orfordness beacon there were previously little data on night errors in direction-finding and none for the rotating beacon. In connection with direction-finding the data appear to be confined to those contained in a previous official report of the Radio Research Board† giving the results of some experiments conducted on a wavelength of 1 000 m between the National Physical Laboratory, Teddington, and the Orford direction-finding station. This work only covered two or three days' experiments, however, and it was considered desirable to carry out a few tests on the subject with the Orfordness beacon.

The results of a series of tests made at three receiving stations between November 1929 and April 1930 are given in Table 4. The second and third stations were about 3 and 10 miles respectively from Teddington, and their distances and true bearings from Orfordness are as given in Table 3. The manner in which the bearings vary at night, as distinct from their steadiness during the daytime, is illustrated by Fig. 2, which shows four records taken with an automatic apparatus described in another paper.‡ In Fig. 2, records (*a*) and (*b*) were taken in the daytime and it will be noted that the successive signal minima agree with each other within about 1°. In records (*c*) and (*d*), however, taken at night, it is seen that the position of the minima varies over some 10° to 26°. In some cases the signal minima were so flat that the recorder failed to operate, while in one case three minima are seen to have occurred during one revolution of the beacon.

Most of the observations in Table 4 were made during 2-hour periods beginning at midnight, although in two or three cases the periods of observation were in the neighbourhood of sunset. The tabulated summary of results shows that at this range of 93 miles overland the error in observed bearing at night may range up to as much as 21°, while any number up to 60 per cent of the observations made may be in error by more than 5°. All these readings were made by skilled observers who have had several years' experience in this work and who, as indicated above, normally obtain 100 per cent of observed bearings correct within 2° in the daytime. In view of this fact, the serious nature of the night errors is shown by the low proportion of night observations which come within this limit. It is noteworthy that during certain periods in January and February the magnitude of the errors was less than those experienced both before and after this time, a fact which, it is believed, is consistent with the results of other experiments on wave phenomena and which is probably due to a decrease in the reflection coefficient of the Heaviside layer during this time. In addition to the variation in apparent bearings observed at night, it is a common experience both in direction-finding and in working with the rotating beacon that the signal minimum becomes blurred or flat due to interference effects. The extent to which this makes a bearing observation difficult or impossible depends upon the skill of the observer, but in the tests under discussion it was decided that where no change in signal intensity was apparent for more than 4 seconds at the minimum it should be classed as " too flat for observation," for it is considered that, however urgently even a very rough bearing may be required, an observation made under such conditions would be valueless in any practical case. From Table 4 it is seen that in some periods practically complete freedom was obtained from such flat signal minima, while at other times it occurred in as many as 50 per cent of the cases of possible bearings.

Previous experience in direction-finding on other wavelengths and in other circumstances has indicated results of a precisely similar nature to those recorded above, and it has further been shown that the phenomena of night errors and flat minima are obtained when the path of transmission is entirely oversea and the range in excess of about 100 miles. This is consistent with some of the results received from ships which were discussed in Section 2 (*e*).

(4) Conclusions.

From the discussion of the results summarized in Section (3) of the paper, it is to be concluded that the inauguration of the rotating-loop beacon at Orfordness has achieved a large measure of success among those for whom it was intended, viz. the Mercantile Marine. The beacon has shown itself to be a most useful wireless aid to navigation and is capable of providing bearings to ships at sea to an accuracy of the order of 2°, when the observers have had some little practice and experience at the work. There is no evidence of any error greater than about 0·5° associated with the beacon installation itself: in directions parallel to the coast line at Orfordness, however, there is the possibility of a deviation of the bearing of the order of 2°. Otherwise at moderate ranges the accuracy is limited entirely by the skill of the observer and the type of watch or clock employed. With the ordinary type of ship's 2-valve receiver the reliability of bearings to the 2° accuracy stated above is obtainable at ranges of 70 to 100 miles at all times, and at ranges up to about 250 miles in the daytime when conditions in regard to freedom from interference are favourable. During the hours of darkness, however, bearings taken at ranges exceeding about 100 miles are subject to certain difficulties in the way of flat and wandering minima. A separate investigation of this effect at fixed stations shows that the errors resulting from observations made under such conditions may be very serious from a navigational point of view. Previous

* See Reference (2). † *Ibid.*, (5). ‡ *Ibid.*, (6).

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experience in the field of direction-finding* has indicated the means by which this limitation may be removed from the rotating beacon itself, and it is probable that, with the continued development of the system, future rotating beacons will have a reliable working range to the average ship's receiver of anything from 250 to 500 miles.

The work described in this paper was carried out for the Radio Research Board, and the author is indebted to the Department of Scientific and Industrial Research for granting permission for its publication. He desires also to acknowledge the assistance rendered by Messrs. H. A. Thomas, M.Sc., Associate Member, E. L. Hatcher, C. W. Spencer and H. M. Bristow in connection with the observations given in Section (3) of the paper.

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* See Reference (7).

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MEETINGS OF THE WIRELESS SECTION.

87TH MEETING OF THE WIRELESS SECTION, 7TH JANUARY, 1931.

Mr. C. E. Rickard, O.B.E., Chairman of the Section, took the chair at 6 p.m.

The minutes of the meeting of the Wireless Section held on the 3rd December, 1930, were taken as read and were confirmed and signed.

A paper by Messrs. F. M. Colebrook, B.Sc., and R. M. Wilmotte, M.A., Associate Member, entitled "A New Method of Measurement of Resistance and Reactance at Radio Frequencies" (see page 49), and a paper by Mr. E. B. Moullin, M.A., Associate Member, entitled "A Variable-Capacitance Cylindrical Condenser for Precision Measurements, and a Wavemeter for Short Wavelengths" (see page 59), were read and discussed.

The meeting terminated at 7.45 p.m. with a vote of thanks to the authors, which was moved by the Chairman and carried with acclamation.

88TH MEETING OF THE WIRELESS SECTION, 4TH FEBRUARY, 1931.

Mr. C. E. Rickard, O.B.E., Chairman of the Section, took the chair at 6 p.m.

The minutes of the meeting of the Wireless Section held on the 7th January, 1931, were taken as read and were confirmed and signed.

A paper by Mr. P. K. Turner, Member, entitled "Some Measurements on a Loud-Speaker in Vacuo" (see page 75), was read and discussed.

The meeting terminated at 7.55 p.m. with a vote of thanks to the author, which was moved by the Chairman and carried with acclamation.

89TH MEETING OF THE WIRELESS SECTION, 4TH MARCH, 1931.

Mr. C. E. Rickard, O.B.E., Chairman of the Section, took the chair at 6 p.m.

The minutes of the meeting of the Wireless Section held on the 4th February, 1931, were taken as read and were confirmed and signed.

A paper by Mr. C. E. Horton, M.A., entitled "The Practical Correction of a Wireless Direction-Finder for Deviations due to the Metal Work of a Ship" (see page 107), was read and discussed.

The meeting terminated at 7.35 p.m. with a vote of thanks to the author, which was moved by the Chairman and carried with acclamation.

90TH MEETING OF THE WIRELESS SECTION, 1ST APRIL, 1931.

Mr. C. E. Rickard, O.B.E., Chairman of the Section, took the chair at 6 p.m.

The minutes of the meeting of the Wireless Section held on the 4th March, 1931, were taken as read and were confirmed and signed.

Papers by Mr. R. Naismith, Associate Member, entitled "Short-Distance Observations on Long-Wave Phenomena" (see page 121) and "Field-Strength Measurements on Daventry 5XX" (see page 127) respectively, were read and discussed.

The meeting terminated at 7.42 p.m. with a vote of thanks to the author, which was moved by the Chairman and carried with acclamation.

91ST MEETING OF THE WIRELESS SECTION, 6TH MAY, 1931.

Mr. C. E. Rickard, O.B.E., Chairman of the Section, took the chair at 6 p.m.

The minutes of the meeting of the Wireless Section held on the 1st April, 1931, were taken as read and were confirmed and signed.

The Chairman announced that the following nominations had been made by the Wireless Section Committee to fill the vacancies which would occur on the Committee on the 30th September, 1931:—

Chairman: Colonel A. S. Angwin, D.S.O., M.C., B.Sc.

Vice-Chairman: L. B. Turner, M.A.

Members of Committee: H. Faulkner, B.Sc., Prof. J. T. MacGregor-Morris, W. J. Picken and T. Wadsworth.

In the event of a ballot for the new Committee being required, Messrs. E. W. Moss and F. Tandy were appointed scrutineers.

An informal discussion, opened by Mr. C. O. Browne, B.Sc., took place on the subject of "Technical Problems in connection with Television."

The meeting terminated at 7.30 p.m. with a vote of thanks, moved by the Chairman, to Mr. Browne for his introductory remarks.

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EXPLANATION OF ABBREVIATIONS.

(P) indicates a reference to the general title or subject of a paper or address.
(p) indicates a reference to a subject dealt with in a paper or address of which the title is not quoted.
(D) indicates a reference to a discussion upon a paper or address of which the general title or subject is quoted.
(d) indicates a reference to a subject dealt with in a discussion on a paper or address of which the general title is not quoted.

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